THE DESIGN OF AN 885-MEGACYCLE TELEVISION TRANSMITTER

by

KENNETH EUGENE FULTZ

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INTRODUCTION

Object

The object of this thesis was to develop the design of an 885-megacycle television transmitter for experimental use by station KA2XBD, which is an experimental television station owned and operated by Kansas State College.

Transmitter Requirements

It was essential that the transmitter have a band pass of 5 megacycles so that standard 525-line television and its associated sound could be broadcast over the same transmitter. This is to be accomplished by frequency modulating a 4.5-megacycle subcarrier and mixing it with the video signal before it is applied to the modulator of the transmitter. The power output required was low because the primary use for the transmitter will be to obtain propagation data at 885 megacycles. However, the antenna was designed for high gain and a desirable broadcast pattern. The transmitter design included the use of as much surplus equipment as possible. This policy was in keeping with the overall object of station KA2XBD to investigate the cost of providing television to small population centers.

Scope

Included in this thesis are the design calculations for the final amplifier, modulator, exciter, and the development of the antenna. Also included are discussions of the tubes and cir-

cuits employed in the transmitter. This thesis does not include the video amplifiers preceding the modulator tube.

Method of Attack

The design of the 885-megacycle transmitter was approached from both a theoretical and an experimental viewpoint. A complete layout for the transmitter was first developed in a qualitative manner. The tubes and circuits for this layout were chosen in accordance with the principles to be discussed later. It should be pointed out here that this layout incorporated the use of the more conventional type of circuits extended into the ultra-high-frequency region instead of the more costly and difficult to construct microwave structures. After the proposed transmitter layout was completed, the radio-frequency stages of the transmitter were constructed and the parameters affecting their operation determined. With this additional information about the circuits employed, the final transmitter design was made.

PLATE-MODULATED FINAL AMPLIFIER

Choice of Tubes for Final Amplifier

At the present time there are not many electron tubes which will operate satisfactorily in the 1000-megacycle region. Klystrons, magnetrons, and specially designed space-charge control tubes are the principal types of tubes useful at this frequency. Klystrons are more readily adapted to oscillators than amplifiers, and a satisfactory method of modulating magnetrons has

not been developed. This eliminates all of the principal types except space-charge control tubes. The main class of tubes within this type are the disk-seal triodes popularly called "lighthouse tubes". Because of their physical construction, they lend themselves readily to use with ultra-high-frequency tank circuits. Since one of the requirements of this project was to make an 885-megacycle transmitter as economically as possible and high power output was not too important, the larger and more costly lighthouse tubes were not chosen. Instead the 2C43 lighthouse tube was selected because it was available on the surplus market. The mechanical construction of a typical lighthouse tube is shown in Plate I.

Choice of Circuit for Final Amplifier

Push-pull operation of two 2C43 tubes in the final amplifier was decided upon since it gives greater power output and simplifies tank circuit design. Push-pull operation also has the advantage at very high frequencies of placing the tube interelectrode capacitances in series, thus reducing the capacitive loading on the tank circuits. Two 2C43 lighthouse tubes operating in push-pull will furnish a power output of 18.3 watts at the crest of the modulation cycle.[1]

Grid separation or grounded grid construction was chosen because it does not require neutralization. Lighthouse tubes are well suited for this type of circuit because the plate and

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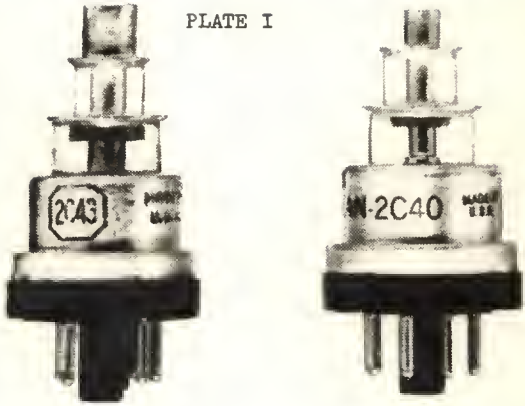
[1] For final amplifier design calculations see Appendix page 75.

EXPLANATION OF PLATE I

Fig. 1. Photograph of two typical lighthouse tubes.

Fig. 2. Mechanical details of the 2C40 lighthouse triode.

PLATE I



(a) (b)

FIG. 1·1.—Typical lighthouse tubes: (*a*) the 2C43 and (*b*) the 2C40.

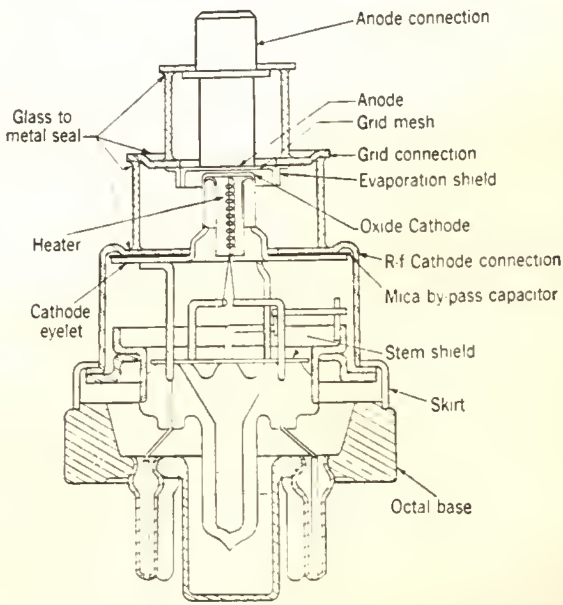


FIG. 1·2.—Mechanical details of the 2C40 tube.

cathode connections are on different sides of the grid plane. Also the plate-to-cathode capacitance is small. This allows complete separation of the cathode and plate tank circuits when the grid plane lies in the plane of the shield separating the two tanks. The final amplifier constructed in accordance with this principle showed no tendency to oscillate.

Three possible methods of providing bias for the final amplifier were considered--cathode bias, grid-leak bias by floating the power supply off of ground, and grid-leak bias with the grids connected to ground through a condenser. These methods of obtaining grid bias are shown in schematic form in Plate II. Grid-leak biasing by floating the power supply off ground was discarded because it would place a large capacitance across the output of the modulator. Grid-leak biasing with the grids grounded through a condenser was chosen over cathode biasing. Grid-leak biasing was preferred because it prevents the plate voltage swinging below the maximum grid voltage during the trough of the modulation cycle.[1] To accomplish this type of biasing without loss of plate and cathode tank separation, the grid terminals of the tubes are connectod to a brass plate which is separated from the center partition of the shield by a sheet of mica. These two plates with mica dielectric form the condenser which maintains the grids at radio-frequency ground.

Lighthouse tubes have been primarily used with coaxial line

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[1] Frederick E. Terman, Radio Engineering, Third Edition (New York: McGraw-Hill), 1947, p. 471.

EXPLANATION OF PLATE II

Fig. 3. Push-pull final amplifier with grid-leak bias, grids connected to ground at radio frequencies by condensers.

Fig. 4. Grounded grid, push-pull final amplifier with cathode bias.

Fig. 5. Grounded grid, push-pull final amplifier with grid leak bias obtained by floating with power supply off ground.

(In Figs. 3, 4, and 5, the plate and cathode tanks composed of resonant sections of parallel wire transmission line are replaced by their lumped constant equivalents for clarity.)

PLATE II

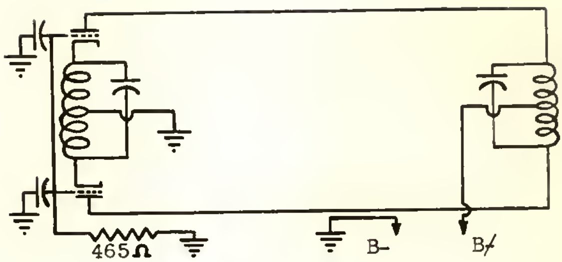


Fig. 3.

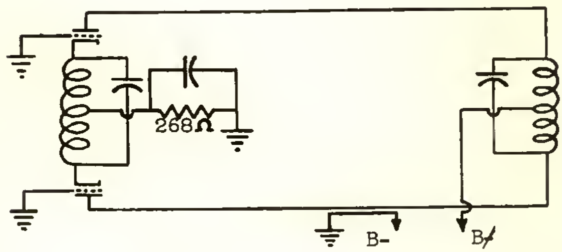


Fig. 4.

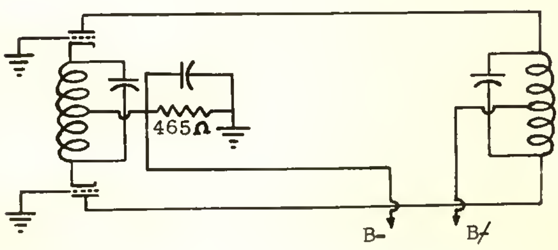


Fig. 5.

resonators. However, coaxial lines do not lend themselves readily to push-pull operation because of the mechanical difficulties in tank construction. If a coaxial tank woro used it would be necessary to have a sliding trombone to vary the length of the line for tuning. An oscillator employing this type of tank was developed as part of the type AN/TPS-8 radar warning equipment.[1] To eliminate the difficult problem of constructing a coaxial tank with its sliding contacts, the plate and cathode tanks for the final amplifier are shorted sections of parallel wire transmission lines.

Experimental Final Amplifier

The original model of the final amplifier was constructed in a brass trough 3 inches deep, 5 inches wide, and 15 inches long. The plate line was made from 3/8-inch silver-plated brass tubing spaced 2 inches between centers. This gives a b/a ratio of 10.7 which exceeds the b/a ratio of 8 for maximum input impedance for a shorted open-wire line.[2] However, tho input impedance is 98 per cent of what it would be for optimum spacing. Tubing 3/8 inch in diameter was chosen because it matches the diameter of the plate cap of the 2C43 and a spacing of 2 inches was selected because the tubes could not physically be placed

[1] Howard C. Lawerence, "High Powor U.H.F. Transmitter," Radio-Electronic Engineering Edition of Radio News, 39:3, May, 1948.

[2] Frederick E. Terman, "Rosonant Lines in Radio Circuits," Electrical Engineering, 53:1046, July, 1934.

closer. The cathode line was constructed of 1 1/2-inch silver-plated brass tubing also spaced 2 inches between centers. The characteristic impedance of the plate line was 272 ohms and the characteristic impedance of the cathode line was 61 ohms. One side of the shield was left open to facilitate movement of the shorting bars and placement of the coupling loops. Cathode biasing was used in the experimental model to simplify the mechanical construction.

Considerable time was spent in attempting to tune the tank circuits to resonance with the excitation being supplied from a variable frequency oscillator capable of tuning to 885 megacycles. It was found that a slight indication of resonance of the cathode tank could be detected on the basis of maximum grid current when the shorting bar was placed very close to the tubes. The plate tank also showed signs of approaching resonance when the two plates were shorted together. This indicated that it would be impossible to tune the tank circuits to the first quarter wave mode. Next, the shorting bars were moved back and a search made for the three-quarter wave mode. More grid current was drawn when the cathode tank was lengthened because better coupling to the oscillator was obtained. However, it was noted that movement of the cathode shorting bar had no appreciable effect upon the grid current. Also adjustment of the plate tank gave no indication of resonance. A check of the standing waves on the cathode line with a neon bulb gave no indication of a voltage maximum on the line. But the neon bulb did indicate the presence of radio-frequency voltage on the inside of the shield. Closer examination showed that the oscil-

lator was exciting the region between the cathode line and the shield instead of between the two conductors of the line. This caused the two tubes to operate in parallel. Reduction in the size of the coupling loop and closer coupling to the cathode line caused push-pull operation of the tubes. A small coupling loop was soldered to a dial light bulb and placed at a current maximum on the cathode line to give a continuous check for push-pull operation while the plate tank was being tuned. Adjustment of the plate tank to resonance gave a very small dip in plate current. The presence of several spurious voltages on the shield were noted, but loading of the amplifier eliminated these. The amplifier was loaded by a 6-volt bulb in series with the coupling loop and a variable condenser. When the output loop was tuned to series resonance, visible power output could be observed by the brilliance of the bulb. However, the power output that could be obtained did not approach the expected power output from design calculations. It was apparent that the losses from the open side of the shield were excessive.

The open side of the shield was covered and variable loading condensers placed across the tank circuits so that tuning could be accomplished from outside of the shield. The coupling loops were reduced in size and placed closer to the lines. This amplifier performed very satisfactorily. The increased shielding reduced the losses and the improved coupling loops eliminated the tendency to excite spurious modes of operation. Power output of 12 watts was observed and it would have been more except that the oscillator being used as an exciter was not capable of furnishing

sufficient driving power. A photograph of the experimental amplifier is shown in Plate III.

Final Amplifier Design Summary

The final amplifier was left unchanged in the final design except for the use of grid-leak bias. The effect of tuning the tanks to the three-quarter wave mode was to reduce the shunt impedance of the tank circuits. This was not too objectionable because the unloaded shunt impedance of the plate tank is 1.25×10^5 ohms and the required loaded tank impedance is 27,000 ohms. The power output of the final amplifier from design calculations is 18.3 watts at the crest of the modulation cycle and the required driving power is 2.67 watts.

MODULATOR

Method of Modulating Final Amplifier

The choice of triodes for the final amplifier ruled out the possibility of using screen-grid modulation. This left only plate modulation and grid modulation as possibilities among the more generally used methods of modulating class C amplifiers. Another possibility was to modulate the driver and operate the final amplifier as a linear class B radio-frequency amplifier. However, the effects of grid modulation and modulation of the driver upon the plate circuit of the final amplifier are essentially the same and the relative merits of plate modulation versus grid modulation only will be discussed.

The lower plate efficiency resulting from grid modulation

EXPLANATION OF PLATE III

Photograph of the experimental final amplifier with the cover turned back. The light bulbs used for load may be seen below the plate line.

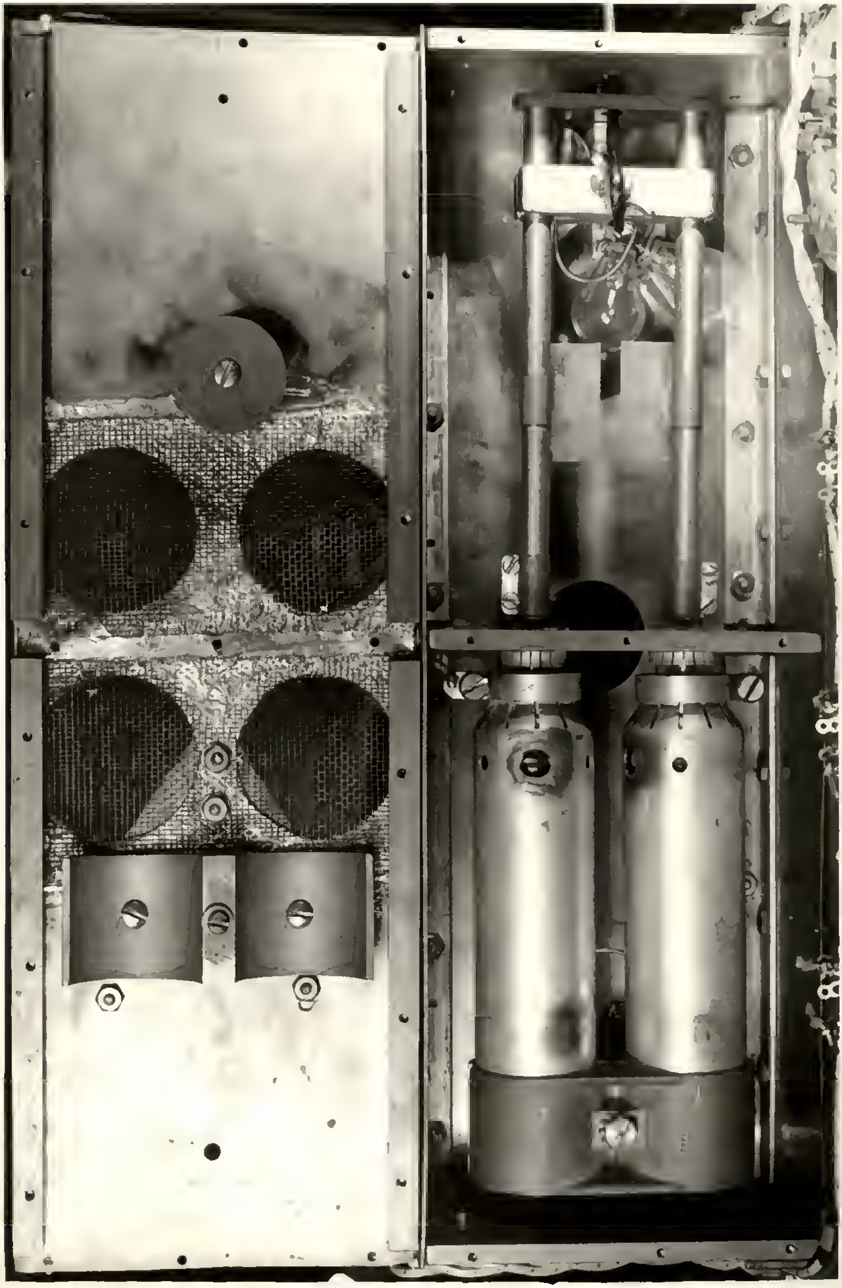


PLATE III

The lower plate efficiency resulting from grid modulation than from plate modulation was not of primary importance here. This is because the plate dissipation at the crest of the modulation cycle as shown by the design calculations is less than one-half of the maximum rated plato dissipation. In other words, the power output of the final amplifier is limited by the maximum cathode emission and not by the maximum plate dissipation.

The main disadvantage of plate modulation is that it requires large modulating power. However, the linearity of modulation is greater with the plate-modulated amplifier than with the grid-modulated amplifier and the plate-modulated amplifier is less critical in adjustment. There is a tendency for the positive peaks of the grid-modulated wave to be flattened because of imperfect regulation of the radio-frequency exciting stage. Also the positive peaks are reduced due to loading of the modulator output when the grid current of the class C amplifier becomes large. The large modulator power required for plate modulation was not considered a serious disadvantage because of the low power output of the final amplifier. At the crest of the modulation cycle of an amplitude-modulated wave, the peak power in the side bands is one-half of the crest power output of the modulated final amplifier (average side band power 1/8 of the effective power output of final at crest conditions). For a crest power output of 18.3 watts, the peak powsr in both side bands is 9.15 watts. For plate modulation, the modulator must supply this peak power divided by the final amplifier efficiency. From design calculations the efficiency of the final

amplifier is 57.1 per cent. The peak modulator power required to modulate the final amplifier 100 per cent is only 16 watts. In view of these facts, plate modulation of the final amplifier was chosen.

The transmitter was designed for transmission of a standard video signal. Consequently, the modulator had to be capable of modulating the radio-frequency output of the final amplifier with signals varying in frequency from 30 cycles per second to 5 megacycles per second. This immediately eliminated the possibility of transformer coupling the modulator to the final amplifier. The other alternative was direct or impedance coupling. Since inductive coupling was not desirable because of the great variation in the impedance of the coupling element over the band pass of the modulator, resistance coupling of the modulator to the final amplifier was selected.

For a class C amplifier properly adjusted for distortionless modulation, the relation between the d-c plate voltage and the d-c plate current is linear. For this condition the final amplifier presents a constant impedance which is a pure resistance equal to E_{bb}/I_b to the modulator. The principle of direct coupling is to control the voltage drop in the coupling impedance which subtracts from the plate supply voltage and gives a voltage across the final amplifier which varies directly with the plate-to-cathode voltage of the modulator tube. Plate IV shows a circuit diagram of the modulator connected to the final amplifier, the simplified circuit diagram, and the equivalent circuit. The a-c load resistance on the modulator is the parallel combi-

EXPLANATION OF PLATE IV

Fig. 6. Resistance-coupled plate modulator circuit diagram.

T_1 = T_2 = 2C43 lighthouse triodes
T_3 = 715A (W.E. 5D21) pulse triods
T_4 = 2A3 power triode
T_5 = 6SJ7 pentode
T_6 = VR 90

C_1 = C_2 = mica condenser formed by mechanical construction
C_3 = C_4 = 100-micromicrofarad condenser built into the 2C43 lighthouse triode
C_5 = 20-microfarad electrolytic, by-passed with mica
C_6 = 0.05-microfarad paper, by-passed with mica

L_1 = 26 microhenries, air-core, shunt-peaking inductance

R_1 = 465 ohms, ½-watt
R_2 = 50,000 ohms, ½-watt
R_3 = 35,000 ohms, 4-watt
R_4 = 15,000 ohms, 2-watt
R_5 = 500,000-ohm potentiometer
R_d = 1,400 ohms, 10-watt
R_A = E_b/I_b = 6,340 ohms
R_L = 2,200 ohms, 110 watts

E_{bb} = 830 volts
E_d = 100 volts

Fig. 7. Simplified modulator circuit. Class C final amplifier replaced by its equivalent input resistance.

Fig. 8. Equivalent circuit of modulator.

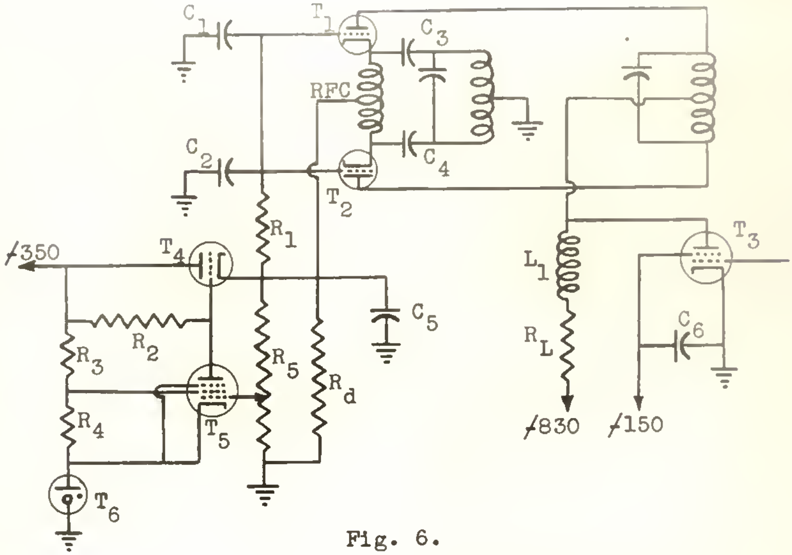


Fig. 6.

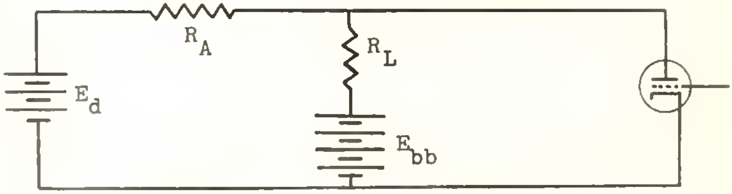


Fig. 7.

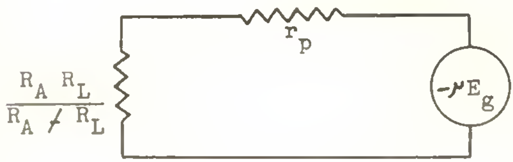


Fig. 8.

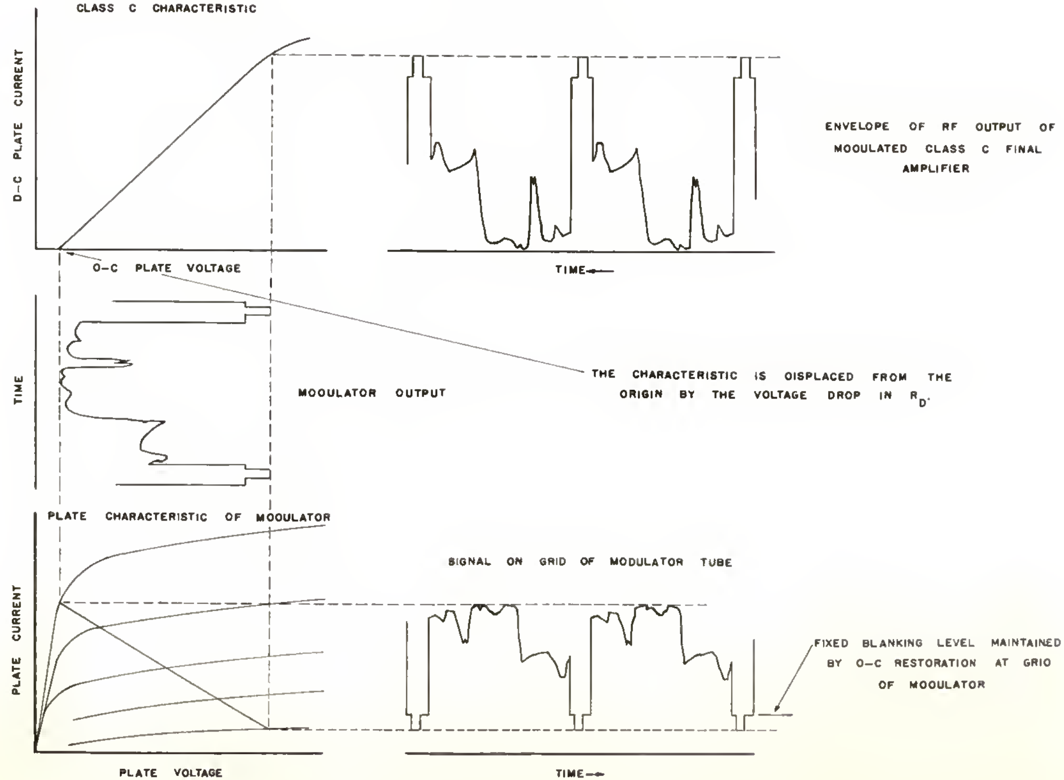
nation of the input resistance of the final amplifier R_A, and the load resistance R_L. The plate voltage on the modulator tube does not go to zero; consequently, the plate voltage on the final amplifier can be reduced to zero for 100 per cent modulation only if the voltage across the final amplifier is less than the voltage across the modulator by a fixed amount. Conventional direct-coupled modulators employ a dropping resistance by-passed with a condenser in series with the plates of the final amplifier. This principle cannot be used for video signals because the average plate current of the final amplifier is not constant. A dropping resistor by-passed by a condenser would tend to remove the d-c component of the video signal. Therefore, a fixed voltage that does not depend upon the amount of the plate current of the final amplifier must be inserted in the circuit to maintain the voltage across the final amplifier below the voltage across the modulator tube by a constant amount. This could be accomplished by a floating power supply in series with the final amplifier plates, but this would increase the shunt capacitance across the output of the modulator. Since the 2C43 lighthouse tubes are provided with both an r-f and a d-c cathode connection, it is preferable to float the cathodes off of ground. The cathode tank may still be grounded and the capacitance in the circuit will not be increased. The cathodes of the final amplifier tubes are maintained at a constant voltage above ground by the dropping resistor R_d. The voltage regulator across R_d keeps the current in R_d constant.

Plate V illustrates the production of the standard modulated

EXPLANATION OF PLATE V

Illustration of the production of the standard amplitude-modulated television radio-frequency output by a resistance-coupled plate modulator having a negative video signal applied to its grid.

PLATE V


CLASS C CHARACTERISTIC
D-C PLATE CURRENT
O-C PLATE VOLTAGE
TIME
ENVELOPE OF RF OUTPUT OF MOOULATED CLASS C FINAL AMPLIFIER
TIME
MOOULATOR OUTPUT
THE CHARACTERISTIC IS OISPLACED FROM THE ORIGIN BY THE VOLTAGE DROP IN R_D.
PLATE CHARACTERISTIC OF MOOULATOR
PLATE CURRENT
SIGNAL ON GRID OF MODULATOR TUBE
FIXED BLANKING LEVEL MAINTAINED BY O-C RESTORATION AT GRIO OF MOOULATOR
PLATE VOLTAGE
TIME

television output when a negative video signal is applied to the grid of the modulator tube. The signal applied to the grid of the modulator tube is shown after the blanking level has been set by d-c reinsertion. This signal is applied to the plate characteristics of the modulator tube on which the a-c load line has been drawn. The application of the output voltage of the modulator to the class C characteristic of the final amplifier is shown.

Choice of Modulator Tube

The alternating-current load resistance on the modulator tube is $(E_{max} - E_{min})/(I_{max} - I_{min})$. $(E_{max} - E_{min})$ is equal to E_{Amax}, the maximum d-c voltage applied to the final amplifier. Consequently, in order for the load resistance on the modulator to be small, which is necessary for good high-frequency response, $(I_{max} - I_{min})$ must be large. Therefore, the modulator tube must be capable of passing large currents and have a large plate dissipation. It is also necessary that the modulator tube have low output capacitance and it is desirable for its input capacitance to be low.

The capacitance requirements for the modulator tube cannot be met satisfactorily by triodes, and pentodes do not meet the current requirements. Consequently, the modulator tube had to be selected from the tetrode or beam-power classes. The 6L6 beam-power tube was considered, but for the allowable plate dissipation the swing in plate current was too small to give good high-frequency response. A 715A (W.E. 5D21) was selected as the

modulator tube because of its higher current capacity and larger plate dissipation.

Modulator Design Summary

The modulator was not constructed because the primary limiting factor in its operation is the shunt capacitance across the output of the modulator tube. This capacitance depends upon the actual mechanical arrangement employed. If the stray wiring capacitance is kept below 8 micromicrofarads, the total shunt capacitance will be under 20 micromicrofarads and the frequency response flat to 5 megacycles (down 3 db at 5.5 mc.) when shunt peaking is employed. From the modulator design calculations the operating grid bias on the modulator tube is 17.5 volts and the peak-to-peak value of the applied signal 35 volts.[1] The design value of load resistance is 2,200 ohms and the shunt-peaking inductance required is 26 microhenries. For this operating condition the plate dissipation of the modulator tube is 60.7 watts, the power supply voltage is 830 volts, and the harmonic distortion is approximately 5 per cent.

To compensate for variations in modulator tube characteristics the screen-grid voltage should be adjusted to give 325 milliamperes of plate current when the control grid voltage is 0 and the plate voltage is 100 volts. Then the control grid voltage should be adjusted for 50 milliamperes of plate current with 550 volts on the plate. These two points are the limits of operation along the a-c load line, and this adjustment will fix

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[1] For modulator design calculations, see Appendix page 85.

the load line in the design position. Consequently, the load resistance and power supply voltage obtained in the design calculations will be correct. The only departure from design conditions resulting from this adjustment will be a change in the harmonic distortion, grid bias, and signal voltage required if the characteristics of the modulator tube used differ from those used for the design.

EXCITER

Frequency Stability

The experimental license under which this transmitter is to be operated does not specify a specific frequency tolerance. The license simply provides that the transmitter may be operated in the band between 880 and 890 megacycles. A search of the Federal Communication Commission's reports failed to produce a ruling on the allowable frequency tolerance. Reference Data for Radio Engineers lists the tolerance at these frequencies as 0.75 per cent until further action is taken.[1]

Since it appeared that a very liberal frequency tolerance would satisfy FCC regulations, the frequency tolerance was fixed by the requirements of the receiver. It was desired that a receiver employing high Q input resonators be able to receive signals from the transmitter without the use of automatic frequency control or frequent manual tuning. A frequency stability of

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[1] Federal Telephone and Radio Corporation, Reference Data for Radio Engineers, Third Edition (New York: Knickerbocker), 1949, p. 13.

0.05 per cent will satisfy this requirement. The carrier was placed at the center of the allotted band (885 megacycles). Therefore, the output of the transmitter was limited to 885 mc $\pm$0.4425 mc.

Definite frequency control within the limits established above could only be provided by crystal control. To reduce the amount of frequency multiplication required to a minimum, the original exciter layout included the use of a high-frequency crystal. Overtone crystals with resonant frequencies up to 100 megacycles are now available. A crystal with a resonant frequency of 73.75 megacycles, which with a multiplication of twelve times would give a frequency of 885 megacycles, was used.

Manufacturer's Recommended Oscillator Circuit

The experience with high-frequency crystal oscillators has been that they are difficult to adjust for crystal-controlled oscillation and that the obtainable power output is small. The author felt that the oscillator was a critical part of the transmitter and warranted construction for test purposes. A crystal designed for operation at the seventh overtone and with this overtone calibrated to 73.750 mc $\pm$0.05 per cent was obtained. Also the oscillator circuit recommended by the manufacturer of the crystal was received. The recommended oscillator circuit is shown in Plate VI. The oscillator is fundamentally a multivibrator employing a dual-triode with the feedback controlled by the crystal. The plate tank of the second section of the dual-triode is tuned to the second harmonic of the crystal frequency.

EXPLANATION OF PLATE VI

Fig. 9. Circuit diagram of the crystal manufacturer's recommended oscillator circuit.

$T_1 = T_2 = \frac{1}{2}$ 12AT7 dual triode

$C_1 = C_4$ = 10-micromicrofarad ceramic or 3-17 midget variable
$C_2 = C_5$ = 340-micromicrofarad button mica
$C_3 = C_6$ = 10-micromicrofarad ceramic

$R_1 = R_3$ = 300 ohms, 1/4-watt
R_2 = 50,000 ohms, 1/4-watt

L_1 = inductance required to tune C_1 and tube capacitance to resonance at crystal frequency (73.75 mc)
L_2 = inductance required to tune C_4 and tube capacitance to resonance at the second harmonic of the crystal frequency (147.5 mc)

Fig. 10. Circuit diagram of the Treuke oscillator.

$T_1 = T_2 = \frac{1}{2}$ 12AT7 dual triode

C_1 = 100-micromicrofarad ceramic
C_2 = 3-17-micromicrofarad midget variable
$C_3 = C_6$ = 340-micromicrofarad button mica
C_4 = 10-micromicrofarad ceramic*
C_5 = 1.2-10-micromicrofarad Cardwell Trim-Air Midget

R_1 = 2,500 ohms, 1/4-watt
R_2 = 6,600 ohms, 1/4-watt

L_1 = air-cored inductance adjusted for proper grid drive
L_2 = air-cored inductance adjusted for parallel resonance with capacitance in cathode circuit at crystal frequency (73.75 mc)
L_3 = inductance required to resonate C_2 and tube capacitance at crystal frequency (0.31 microhenry for 73.75 mc)
L_4 = inductance required to resonate C_5 and tube capacitance at the second harmonic of the crystal frequency (0.233 microhenry for 147.5 mc)
L_5 = output coupling loop; capacitive coupling may be used.

* The drive on the doubler section may be adjusted either by varying C_4 to change the voltage division between C_4 and the input impedance of the next stage, or by tapping down on L_3.

PLATE VI

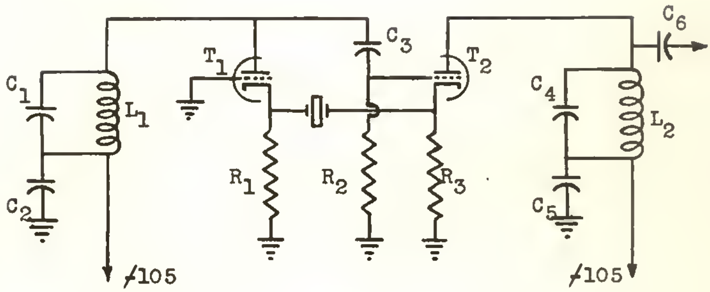


Fig. 9.

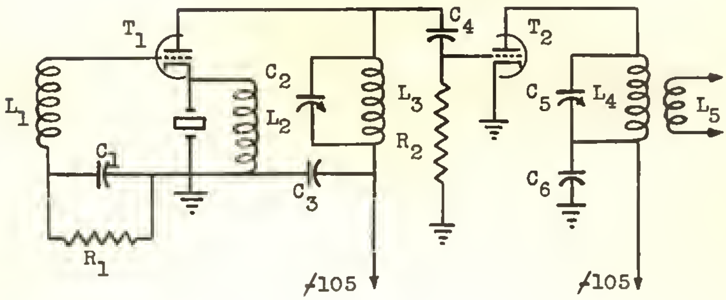


Fig. 10.

The operation of the oscillator may be explained in the following manner. For steady-state conditions the voltage across the plate tank of the first section will be sinusoidal because it is a high Q resonant circuit. The plate-to-ground voltage of section one is also the grid voltage applied to tho second section. The peak value of this grid voltage is sufficient to cause class C operation of the second section. In othor words, the second section will be cut off during the negative half-cycle of grid voltage and rise to a positive peak sufficiently high to draw grid current which flows through the grid resistor to provide bias for the section. The plate current in the second section flows in sharp pulses. These plate current pulses produce a voltage across the cathode resistor. The fundamontal component of this voltage is passed through the crystal and forms the feedback voltago for the first section. The feedback voltage is in phase with the voltage applied to the grid of the second section. Since the foedback voltage is applied to the cathode of the first section, with the grid grounded, the grid-to-cathode voltage is 180 degrees out of phase with the feedback voltago. The plate current flowing in the first section is in phase with the grid voltage applied to the first section and the plate-to-ground voltage is 180 degrees out of phase with the plate current. Consequently, the plate-to-ground voltage of section one is in phase (360 degroes out of phase) with the feodback voltage and in phase with the grid voltago of the second section. Therefore the phase relations are correct for sustained oscillation.

A more quantitative discussion of the operation of the oscillator may be made by considering only the fundamental frequency components of the space currents. The peak value of the fundamental component of the space current of section one will be designated by I_1 and the peak value of the fundamental component of the space current of the second section will be called I_2. Since the two sections conduct maximum currents alternately, these two currents are 180 degrees out of phase. The crystal is a short circuit to fundamental currents and the cathode resistors operate in parallel. (They are in parallel for the fundamental frequency component of current only.) Consequently, they carry currents that are in phase and equal to $(I_2 - I_1)/2$ in magnitude. I_2 is larger than $(I_2 - I_1)/2$, so the current through the crystal must be in phase with I_2 and equal to $(I_2 + I_1)/2$ in magnitude. Therefore the fundamental component of current in the cathode resistor of section one is 180 degrees out of phase with the fundamental component of space current in section one. This condition can exist only if I_1 is less than I_2. The peak value of the feedback voltage is $(I_2 - I_1)R_c/2$. This gives the proper regulatory action to the oscillator. If I_1 decreases, the feedback voltage increases; if I_1 increases, the feedback voltage decreases. Maximum feedback voltage would occur if I_1 were equal to zero. For good class C operation of the second section of the 12AT7, I_2 is 20 milliamperes. The maximum feedback voltage that could exist would be 3 volts for a 300-ohm cathode resistor. Actually the fundamental component of space current in the first section is approximately 4 milliamperes. Then the feedback volt-

age is 2.4 volts and tho tubo operates class AB_2 with a bias of approximately 1.5 volts.

This oscillator was first constructed with slug-tuned coils for the plate tanks. Tho tanks were first tuned to resonance at the desired frequencies with a grid-dip meter. With the crystal in place, an attempt was made to tune the 73.75-megacycle tank for maximum grid current in the second section. However, no maximum could be found, the grid current continued to increase as the slug was removed from the coil, but it did not rise sharply as though approaching a resonant peak. The same result was observed when the doubler tank was tuned; i.e., the output of the oscillator increased as the slug was removed from the coil. Tests with a variable frequency oscillator and a grid-dip meter proved that the slugs were capable of tuning the tanks to the desired frequencies. The coils were removed and their Q determined with a Q-meter. They were found to have very low Q's. Also the Q increased considerably as the slugs were removed from the coil. This indicated that the grid curront in the second section increased as the slug was removed because of decroased losses and not because of resonance of the tank.

The slug-tuned coils were replaced by air-cored coils tuned to resonance by adjusting the number of turns and spacing. The tanks were resonated to the desired frequencies with the tubes in place and filaments heatod but without plate voltage by the use of a grid-dip metor. When plate voltage was applied, the grid current in the second section of the dual-triode and the

power output at 147.5 megacycles were more than doubled. A check for crystal-controlled operation was made with a General Radio frequency meter. The oscillations were not crystal-controlled. A careful search of the dial on the G.R. frequency meter produced beats every 68 kilocycles in the vicinity of 147 megacycles. The oscillator was blocking.

Since the oscillator had been carefully constructed with short leade and the best apparent arrangement of parte, no couree of action which would keep the oscillator from blocking was obvious. Rather than rebuild the oscillator in the hope of finding another arrangement of parts or value of circuit components which would prevent blocking of the oscillator, it was decided to try another oecillator circuit.

Experimental work with harmonic-type crystale at 48 megacycles has been done by Treuke.[1] He recommended the oscillator circuit given in Plate VI. This oscillator is fundamentally a Hartley oscillator with no mutual coupling between the grid and plate inductances. At the crystal frequency the cathode is connected to ground through the series resonant equivalent circuit of the crystal. For frequencies other than the resonant frequency of the crystal, the cathode ie removed from ground by the high impedance of the crystal in parallel with the cathode inductance and the circuit is very degenerative. The cathode inductance also provides the d-c connection to the cathode.

[1] Gus Treuke, "A Regenerative Oscillator for Harmonic-Type Crystals," QST, 33:48, December, 1949.

Advantages of the Treuke Oscillator

One advantage of the Treuke oscillator is that the oscillator section of the dual triode operates class C instead of class AB_2. Therefore the power output obtainable is larger than for the manufacturer's oscillator. A second advantage is that the cathode of the oscillator tube is at radio-frequency ground potential during crystal-controlled operation. Consequently the cathode-to-heater capacitance of the tube does not affect the operation of the oscillator. A third advantage is that the grid excitation can be adjusted by varying the turns in the grid coil.

Adjustment of Treuke Oscillator

The Treuke oscillator was constructed with the cathode of the oscillator tube connected to ground and adjusted for good operation at 73.75 megacycles. It is important to tune the oscillator for a natural frequency very close to the crystal frequency or the crystal will have no effect. Then the capacitance of the crystal was measured at 60 cycles by a capacitance bridge. This capacitance was 10 micromicrofarads for the crystal used. The cathode short was removed and a capacitance equal to that of the crystal at 60 cycles placed across the crystal holder. The cathode inductance was tuned to parallel resonance with this capacitance at the crystal frequency. Thus, with the crystal in place, the cathode circuit becomes a parallel resonant circuit shunted by the series resonant quartz crystal. Therefore the

cathode is at ground potential only at the crystal frequency and the oscillator is highly degenerative either side of this frequency. The value of the cathodo inductance also affects the starting of oscillations. In case oscillations do not start readily, it may be necessary to adjust the grid excitation. As with all regonerative crystal oscillators, the Treuke oscillator must be carefully checked to be sure that the oscillations are crystal-controlled.

Original Exciter Layout

As originally planned, the exciter would consist of a 12AT7 dual-triode as oscillator-doubler, a 6J6 with both sections in parallel as a tripler, and two 368 doorknob tubes as a push-push doubler. This arrangement was selected because no two tanks would be tuned to the same frequency and none of the stages would require neutralization. This arrangement also allowed all plate tanks to be single-ended,and capacitively-loaded, coaxial-line resonators could be used at the higher frequencies.

In the original exciter layout, capacitively-loaded coaxial-line resonators were used for the plate tanks of the 6J6 tripler and the 368 push-push doubler. The frequencies were 442.5 mc and 885 mc, respectively. Resonators of this type were available in type R-89-ARN-5A surplus radar receivers. These resonators were 5-3/4 inches long. The radius of the center conductor was 1/8 inch and the inner radius of the outer cylinder was 1/2 inch. The loading condenser plate was removable to allow coarse tuning by using different sizes of condenser plates.

Fine tuning was accomplished by the cap which screwed onto the outer cylinder and varied the loading capacitance. The high-impedance input to the resonator was provided by a tap on the center conductor. Power was removed from the resonator by a balanced coupling loop. The entire resonator was silver-plated.

Tank circuits of this type have many advantages. They can be used with conventional type tubes. The radiation losses are low because of the shielding inherent in a coaxial structure. The Q is high and the possible tuning range is large. They are physically small at 400 megacycles and more easily used mechanically than parallel wire lines.

If appreciable capacitance is introduced at the end of the line, the resonant frequency is approximately the resonant frequency of that capacitance and the inductance of the shorted coaxial transmission line formed by the center conductor and the outer cylinder. Experimental data showed that the resonator described was capable of resonating the output capacitance of a 6J6 at frequencies as low as 150 megacycles when a loading condenser plate 7/8 inch in diameter was used. If the gap length is large so that there is no appreciable capacitive loading, and the center conductor length is large compared to the inner radius of the outer cylinder, resonance will occur when the center conductor is approximately a quarter wave long. This is a frequency of 515 megacycles for the resonators described. Therefore, it is possible to tune the resonator to 442.5 megacycles by properly adjusting the size of the loading capacitance.

As the length of the center conductor becomes small, the

resonant frequency does not approach infinity, but approaches instead the resonant frequency for the $TM_{0,1,0}$ mode in a cylindrical cavity. This limiting value is the velocity of light divided by 2.61 times the inner radius of the outer cylinder, or 9000 megacycles for the resonator described. Consequently it is possible to tune the line to resonance at 885 megacycles by shortening the center conductor.

Resonators of the capacitively-loaded coaxial-line type were chosen for the original exciter layout because they operate satisfactorily at the intermediate frequencies. The intermediate frequencies constitute the gap between frequencies where lumped constant circuits perform well and frequencies where it is necessary to use microwave tubes and tank structures. However, they do have some serious limitations which will be discussed later.

The exciter shown in schematic form in Plate VII was constructed. The first difficulty experienced with this layout was the determination of the resonant frequency of the coaxial tank circuits. Determination of resonance by watching for a dip in plate current was practically impossible because the resonant peaks were too sharp and because the plate current dip was small for the frequency multiplier stages. Resonance could be passed without apparent indication unless an exceeding amount of care was used. This complication was due in part to the poor contact made by the threads between the cap and cylinder. The most effective method of tuning the tanks was to place the output of a variable-frequency oscillator across the input of the tank when no plate voltage was applied and tune the oscillator

EXPLANATION OF PLATE VII

Circuit diagram of the original exciter. The circuit diagram for the experimental Treuke oscillator is also shown.

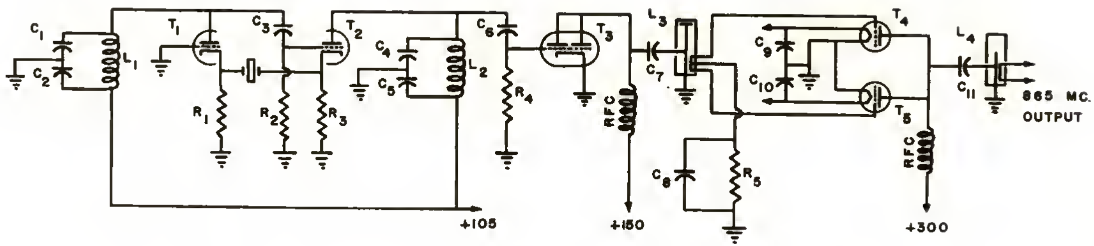
$T_1 = T_2 = T_6 = T_7 = \frac{1}{2}$ 12AT7 dual triode
T_3 = 6J6 dual triode
$T_4 = T_5$ = W.E. 368 doorknob tube

$C_1 = C_4 = C_{16}$ = interelectrode and stray capacitance
$C_2 = C_5 = C_9 = C_{10} = C_{14} = C_{17}$ = 340-micromicrofarad button mica
$C_3 = C_6 = C_8 = C_{15}$ = 50-micromicrofarad ceramic
$C_7 = C_{11} = C_{18}$ = 25-micromicrofarad ceramic
C_{12} = 100-micromicrofarad ceramic
C_{13} = 3-17-micromicrofarad midget variable

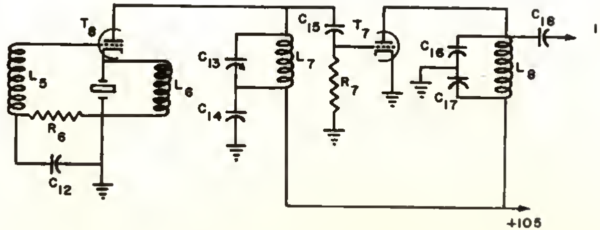
$R_1 = R_3$ = 300 ohms, 1/4-watt
R_2 = 50,000 ohms, 1/4-watt
$R_4 = R_5$ = 5,000 ohms, 1/4-watt
R_6 = 2,500 ohms, 1/4-watt
R_7 = 10,000 ohms, 1/4-watt

L_1 = air-cored inductance adjusted for parallel resonance with C_1 at crystal frequency
L_2 = air-cored inductance adjusted for parallel resonance with C_4 at the second harmonic of the crystal frequency (147.5 mc.)
L_3 = capacitively-loaded coaxial-line resonator, Lapp Insulator Co., Inc., #C222619-12
L_4 = coaxial-line resonator, Lapp Insulator Co., Inc., #C222619-12 with the center conductor shortened
L_5 = air-cored inductance adjusted for proper grid drive
L_6 = air-cored inductance adjusted for parallel resonance with capacitance in cathode circuit at crystal frequency (73.75 mc)
L_7 = inductance required to resonate C_{13} and tube capacitance at the crystal frequency
L_8 = inductance required to resonate the interelectrode capacitance and the stray capacitance at the second harmonic of the crystal frequency (147.5 mc)

PLATE VII



THE ABOVE OSCILLATOR–DOUBLER WAS REPLACED BY THE TREUKE OSCILLATOR BEFORE EXPERIMENTS WITH THE REMAINDER OF THE EXCITER WERE STARTED



for maximum tank output. Then by repeatedly changing the oscillator frequency a small increment in the direction of the desired frequency and tuning the tank to this frequency, the tank was resonated at approximately the desired frequency. Then the oscillator was placed across the input to the stage and adjusted for resonance with the stage acting as a straight through amplifier. When the tank was again resonant at the desired frequency, the low-frequency driving voltage was applied and the tank tuned for proper frequency multiplication.

The second disadvantage of the coaxial tanks was their inability to tune to high enough frequencies. Since the tube did not form a part of the tank, the leads to the tank resonated the output capacitance of the tube before the desired frequency was obtained. This occurred in the 6J6 tripler at about 350 megacycles. The tripler was converted to push-pull operation by connecting the plates to a center-tapped coupling loop added to the resonator. This placed the output capacitances of the two sections of the 6J6 in series instead of parallel. The lumped constant tank of the 12AT7 doubler was replaced by a coaxial resonator to provide balanced output.

The third and most serious objection to the layout was discovered during experimentation with the 6J6 push-pull tripler. The circuit oscillated. At first it appeared to be a parasitic oscillation resulting from poor ground connections. However, this was not the case; the 6J6 was operating as a tuned-plate, tuned-grid oscillator. True, the plate tank of the 12AT7 doubler and the 6J6 tripler were completely shielded by virtue of being coaxial tanks, and they were tuned to different frequencies. But

the output loop of the 12AT7 doubler which was connected to the 6J6 grids was exciting the 12AT7 doubler tank at its third harmonic mode by energy transferred to the grids through the interelectrode capacitances. The conclusion reached from this experience was that coaxial tanks cannot be employed successfully in successive stages unless grounded grid construction is used.

The oscillations in the 6J6 tripler could have been suppressed by using a lumped constant tank for the 12AT7 doubler. However, the layout would still include the use of coaxial tanks for the 6J6 tripler and the 368 push-push doubler. A lumped-constant circuit for either tank would be very inefficient. Consequently the exciter was redesigned.

New Exciter Layout

Three important parameters were determined during the experimentation with the original exciter. First, above 400 megacycles the tube and resonator should form an integral unit. If they do not, resonance of the leads between the tube and tank with the interelectrode capacitances will nullify effects of the tank. Second, careful shielding of each tank circuit is essential. This will eliminate the tendency for oscillation of the stages due to the multiple modes which exist in distributed constant resonators. Satisfactory shielding at ultra-high frequencies can be obtained only in grounded-grid circuits. Third, it is exceedingly difficult to obtain a tank which provides adequate shunt impedance and operates in the fundamental mode with available tubes in the 900-megacycle region without resorting to cavity resonators.

The experimental Treuke oscillator performed satisfactorily and remained unchanged in the new exciter layout. The coaxial tank in the 12AT7 doubler was replacod by a lumped-constant circuit. The 6J6 was changed to a push-push doubler employing a capacitively-loaded, coaxial-line plate tank. However, the coaxial line employed is slightly different from those discussed previously. Also available in the type R-89-ARN-5A receiver are coaxial resonators which have the same radii as those describod but are only 3-5/8 inches long. These resonators are equipped with two taps on the center conductor. A resonator of this typo was preferred because it provided unbalanced output for driving the succeeding stage. The next stage had to be a tripler with its plate tank operating at 885 megacycles.

Lighthouse tubes were chosen for the tripler and the driver because they are the cnly readily available tubes which can be used as an integral part of the tank and still operate grounded grid.

Two types of tank circuits were considered for these ultra-high-frequency stages. They were shorted sections of coaxial line and shorted sections of flat-element transmission line. Some experimental work has been done at the MIT Radiation Laboratory with lighthouse tube amplifiers employing coaxial line tanks.[1] The results obtained with amplifiers of this type were

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[1] Donald R. Hamilton, Julian K. Knipp, and J. B. Horner Kuper, Klystrons and Microwave Triodes (New York: McGraw-Hill), 1948, p. 163.

satisfactory. However, they are difficult to construct because of the sliding contacts necessary for tuning. The use of flat-element transmission lines for the tank circuits offered a much simpler and less costly solution. A citizen's band transmitter designed by Koch has used flat-element transmission lines to an advantage.[1] Another advantage of flat-element transmission lines is that the characteristic impedance can be adjusted instead of being fixod by the tube dimensions as is the case for coaxial lines. The full significance of this advantage was not realized until after the tank circuit design calculations were made. A shorted transmission line with a characteristic impedance of 75 ohms will be 2.07 inches long for resonance in the first quarter wave mode with the output capacitance of a 2C40 at 885 megacycles. The author thought that the ability to alter the characteristic impedance of the flat-element transmission line by adjusting the width and spacing of the elements allowed the length to be varied. This is true; however, at the same time, control of the unloaded shunt impedance of the tank is provided.

A plate tank for the 2C40 tripler composed of a shorted section of flat-element transmission line made from elements 2 inches wide and spaced 3/16 inch apart was first considered. The characteristic impedance of this line was 35.35 ohms. The characteristic impedance was purposely made low to make the line physically long. The line length required for resonance at 885

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[1] Arthur R. Koch, "Citizen's Band Transmitter," Electronics, 22:118, December, 1949.

megacycles was 2.68 inches. In order to place a tuning condenser in the open end of the line it was necessary to tap the tube down on the line. The unloaded shunt impedance of this line at a point 1 1/2 inches from the open end was 27,000 ohms and the required loaded tank impedance for the tripler was 24,200 ohms. The shunt impedance of the tank was much too low for good class C operation of the tripler.

Next the width and spacing of the elements were doubled. This did not change the characteristic impedance of the line nor the length required for resonance. However, the unloaded shunt impedance was doubled because the attenuation is inversely proportional to the spacing. The unloaded impedance of 55,400 ohms gave a tank efficiency of 56 per cent. However, the major objection to this arrangement was the tank width in comparison to its length. With the tube located only slightly over an inch from the short and the line 4 inches wide, the currents will not be flowing in straight parallel lines on the conductors. The departure from the conditions under which the design equations used were derived was so great that the tank could not be expected to behave as calculated.

The tank width could have been reduced to 3 inches, but this would have raised the characteristic impedance and shortened the tank. Instead, the tank width was increased to 6 inches. Then the tank was cut in half lengthwise and placed on each side of the tube. For design purposes the line was still effectively 6 inches wide. This gave a plate tank with both ends shorted and the tube placed at the center of the line. The tuning condensers were placed beside the tube and the objection-

able necessity of tapping the tube down on the line eliminated. This procedure increased the distance between the tube and the short and placed the tuning condensers in a position to assist in making the currents flow in parallel lines along the line.

A 2C40 lighthouse tripler feeding a 2C40 power amplifier with both stages employing shorted flat-element transmission lines for tank circuits were chosen to complete the exciter.

Exciter Design Summary

The new exciter layout discussed above removed the objectionable features of the original exciter. The design calculations for the new exciter are given in the Appendix.[1] The circuit diagram for the exciter is shown in Plate VIII. A summary of the power output and driving power required by the various stages of the exciter is given in Table 1.

Table 1. Exciter summary (design values).

Driving power required		Design power output
	Crystal oscillator at 73.75 mc	0.276 watt
0.088 watt	12AT7 doubler to 147.5 mc	0.314
.136	6J6 push-push doubler to 295 mc	1.29
.862	2C40 tripler to 885 mc	1.50
.596	2C40 power amplifier at 885 mc	4.28
2.67	2C43 push-pull final amplifier	18.3

Both 2C40 lighthouse stages were designed to employ shorted flat-element transmission lines for tanks. The tanks are tuned by variable loading condensers. The cathode line for the 2C40 tripler is composed of elements 2 inches wide spaced 3/8 inch

[1] For exciter design calculations, see Appendix page 93 .

EXPLANATION OF PLATE VIII

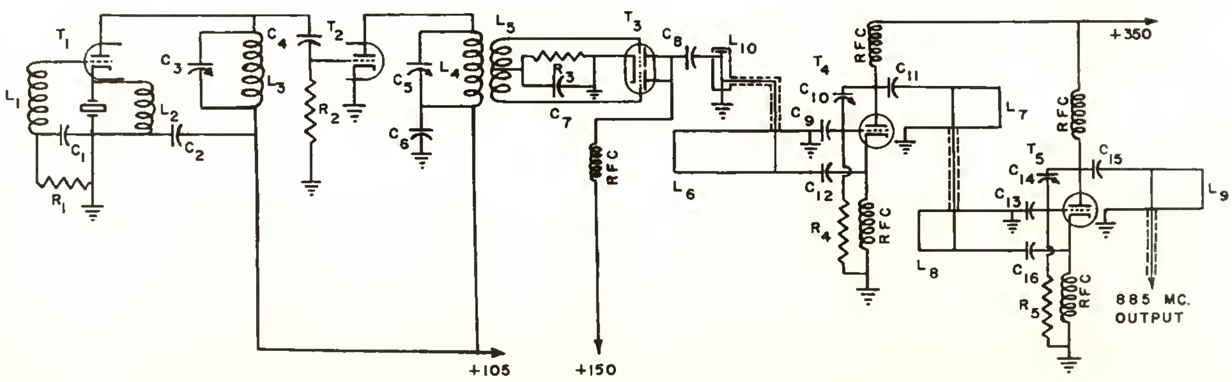
New exciter circuit diagram

T_1 = T_2 = ½ 12AT7 dual triode
T_3 = 6J6 dual triode
T_4 = T_5 = 2C40 lighthouse triode

R_1 = 2,500 ohms, 1/4-watt
R_2 = 6,600 ohms, 1/4-watt
R_3 = 1,560 ohms, 1/4-watt
R_4 = 7,250 ohms, 1/2-watt
R_5 = 1,400 ohms, 1/4-watt

C_1 = 100-micromicrofarad ceramic
C_2 = C_6 = 340-micromicrofarad button mica
C_3 = 3-17-micromicrofarad midget variable
C_4 = 10-micromicrofarad ceramic
C_5 = 1.2-10-micromicrofarad Cardwell Trim-Air Midget, type ZR-10AS
C_7 = 50-micromicrofarad ceramic
C_8 = 25-micromicrofarad ceramic
C_9 = C_{13} = mica condenser formed by mechanical construction
C_{10} = C_{14} = variable tuning condensers for plate lines
C_{11} = C_{15} = mica condenser formed by mechanical construction
C_{12} = C_{16} = mica condenser built into the 2C40 tubes
L_1 = air-cored inductance adjusted for proper grid drive
L_2 = air-cored inductance adjusted for parallel resonance with capacitance in cathode circuit at crystal frequency (73.75 mc)
L_3 = inductance required to resonate C_3 and the tube capacitance at the crystal frequency (0.31 microhenry)
L_4 = inductance required to resonate C_5 and the tube capacitance at the second harmonic of the crystal frequency (0.233 microhenry for 147.5 mc)
L_5 = coupling loop tuned to series resonance with the input capacitance of the 6J6
L_6 = shorted section of flat-element transmission line tuned to resonance at 295 mc
L_7 = shorted section of flat-element transmission line tuned to resonance at 885 mc
L_8 = shorted section of flat-element transmission line tuned to resonance at 885 mc
L_9 = shorted section of flat-element transmission line tuned to resonance at 885 mc
L_{10} = capacitively-loaded, coaxial-line resonator, Lapp Insulator Co., Inc., #D-223920 resonant at 295 mc

PLATE VIII


T1
L1
C1
R1
L2
C2
C3
L3
C4
T2
R2
C5
L4
C6
L5
R3
C7
T3
RFC
C8
L10
L6
T4
C9
C10
C11
C12
R4
RFC
L7
L8
RFC
T5
C13
C14
C15
C16
R5
RFC
L9
+105
+150
+350
885 MC.
OUTPUT

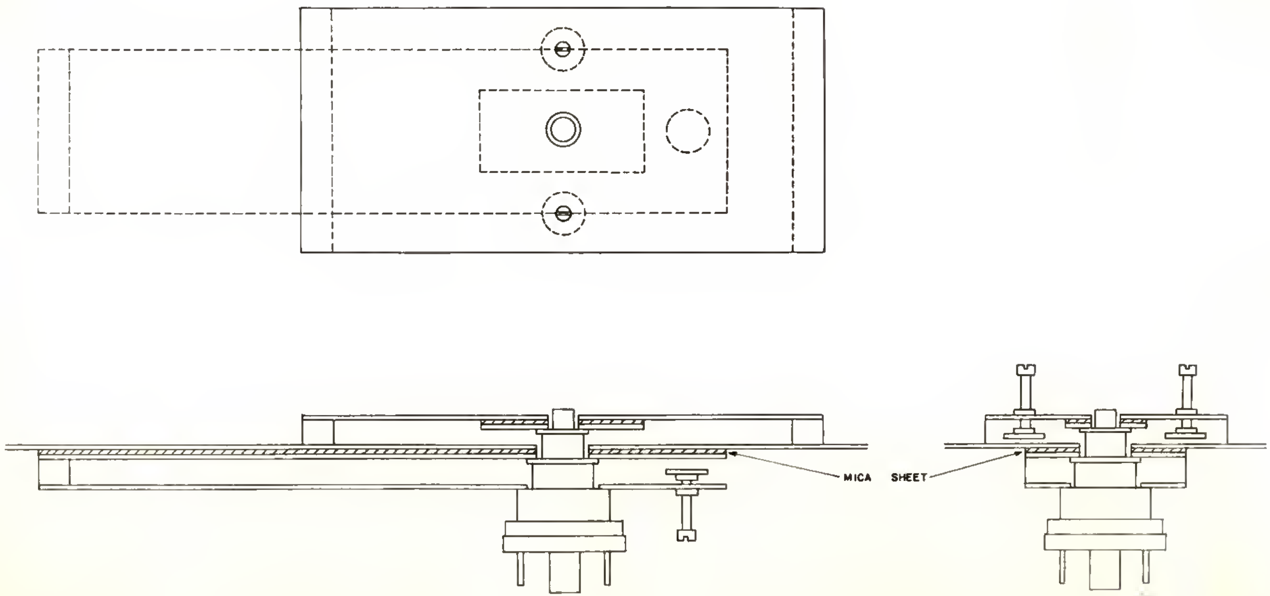
apart. A spacing of 3/8 inch was chosen because the space between terminals of the 2C40 is 3/8 inch. The line is 8.05 inches long and the tube is tapped down 2 inches from the open end. The variable tuning condenser is placed in the open end of the line. The plato tanks of the tripler and power amplifier aro identical. They are made of elements 3 inches wide spaced 3/8 inch apart with both ends shorted and the tube placed at the center. A line width of 3 inches was chosen so that the tuning condensers could be placed beside the tube and not be so close to the glass seals of the tube that excessive dielectric losses would occur in the glass. The effective characteristic impedance of the plate line is 23.6 ohms and the tube is 2.79 inches from the short. The total length of the plate tanks is 5.58 inches. The unloaded shunt resistance of the tank is 1.24×10^5 ohms. The cathode line for the power amplifier is made from elements 3 inches wide and spaced 3/8 inch apart. It is constructed and tuned in the same manner as tho plate tanks. The total length of the cathode tank is 5.18 inches. Plates IX and X show the mechanical tank construction of the two lighthouse stages in the exciter.

Connections to the tanks are to be made with clips which can be moved along the line to obtain the desired coupling. The plate tanks of both lighthouse stages are silver-plated. In this design grounded grid construction is used and grid-leak biasing is employed. The stages must be completely enclosed in shields. Care should be exercised to make certain that the shields are not resonant cavities at the tank frequencies.

EXPLANATION OF PLATE IX

Sketches of the 2C40 tripler showing the physical tank arrangement.

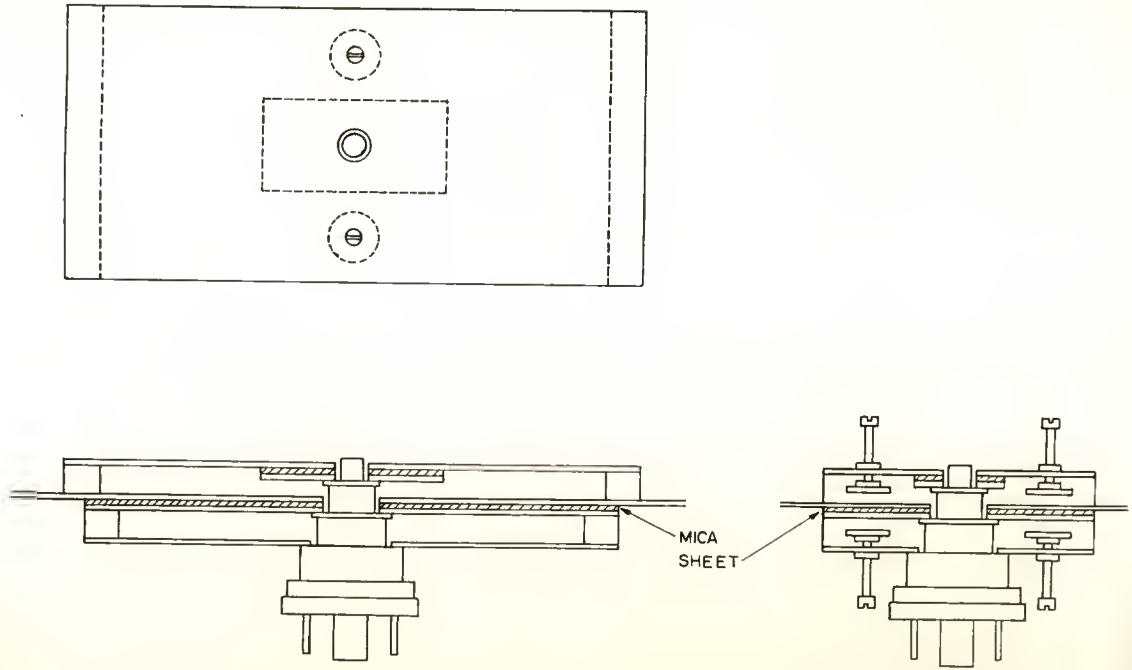
PLATE IX



EXPLANATION OF PLATE X

Sketches of the 2C40 power amplifier showing the physical tank arrangement.

PLATE X



POWER SUPPLY

The power supply was designed to operate the whole transmitter. This eliminated as many sources of trouble as possible during operation of the transmitter. The size of the power supply was fixed by the requirements of the rest of the transmitter. These requirements are listed in Table 2.

Table 2. Power supply requirements.

Transmitter element	Direct current required	
Final amplifier		35.6 ma
Modulator		
Plate	180 ma	
Screen	20	200
Dropping power supply		85
Exciter		
½ 12AT7 oscillator	6.5	
½ 12AT7 doubler	7.2	
6J6 push-push doubler	17	
2C40 tripler	11	
2C40 power amplifier (885 mc)	18.7	60.4
Video amplifiers		90
Total current required		471 ma

To fulfill the power requirements of the various stages of the transmitter, the power supply must have the d-c voltage and current outputs listed in Table 3.

Table 3. Power supply output.

215.6	ma at	830	volts	
225	" "	350	"	regulated
17	" "	150	"	"
14	" "	105	"	"
8	ma for	VR 150		
8	" "	VR 105		
10	" "	bleeder		
497.6	Total			

The current requirements listed above determined the current capacity of the power transformer, the chokes, and the rectifier tubes required. The effective d-c secondary voltage of the power transformer must be 930 volts.

The power supply was a conventional full-wave rectifier with choke input filter. The 830-volt unregulatod output was taken directly from the bleeder. The 350-volt regulated output was obtained with a series pass-tube regulator. The 105-volt and 150-volt outputs were obtained by placing resistors in series with voltage regulator tubes. The circuit diagram of the power supply is given in Plate XI.

ANTENNA

Requirements

To be satisfactory for television usage an antenna must be broad band. This is essential to reduce the changes in input impedance of the antenna with frequency. If the input impedance is not constant over the range of frequencies transmitted, excessive standing waves on the transmission line feeding the antenna will result. In television systems,standing waves in the transmission lines will cause ghost images to appear on the receiving screen and detract from the picture quality. Therefore, the first requirement of the antenna was that it have a broad band characteristic.

The second requirement of the antenna was that it match a coaxial transmission line without the use of impedance transformation. This was desirable because any tuned impedance transformer

EXPLANATION OF PLATE XI

Circuit diagram of the power supply for the transmitter.

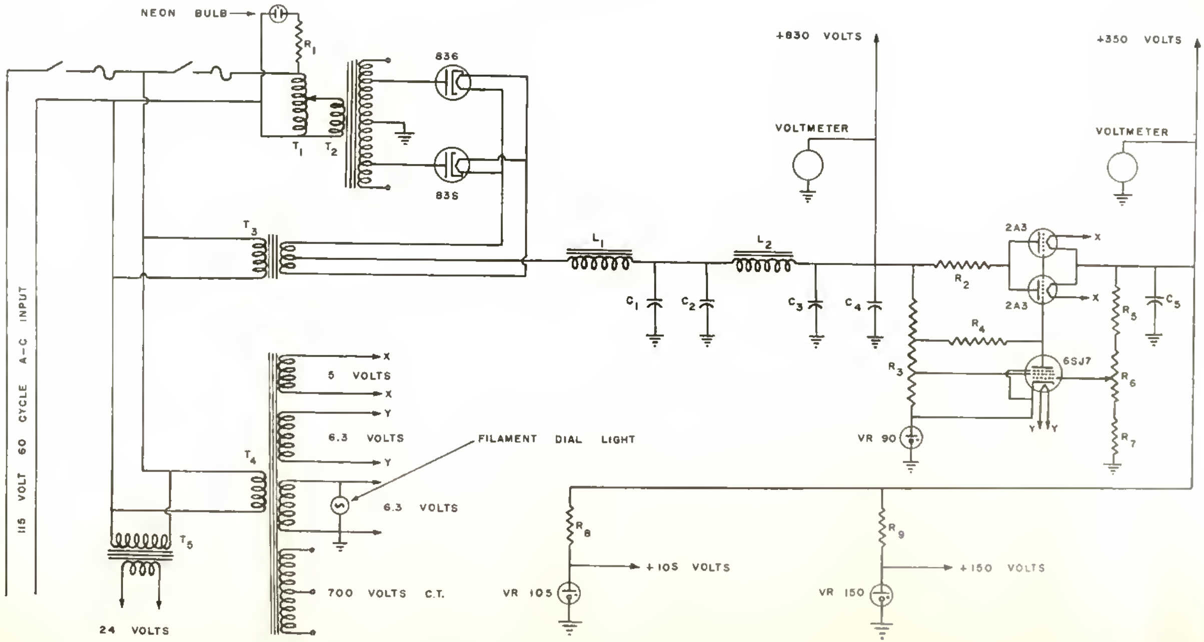
T_1 = 115-v pri., 70/130-v sec., variable output auto transformer
T_2 = 115-v pri., 3120/2530 ct sec. 925-va plate transformer
T_3 = 115-v pri., 2.5-v, 10-a sec. filament transformer
T_4 = 115-v pri., 5-v 2-a, 6.3-v 0.3-a, 6.3-v 7.5-a, 700-v 85-ma sec., power transformer
T_5 = 115-v pri., 24-v 2-a sec. filament transformer

$C_1 = C_2 = C_3 = C_4$ = 4-microfarad, 1000-volt condensers
C_5 = 1-microfarad, 500-volt condenser

$L_1 = L_2$ = 4-henry, 500-ma choke

R_1 = 50,000-ohm, 1-watt
R_2 = 2,750-ohm, 40-watt
R_3 = 75,000-ohm, 50-watt
R_4 = 50,000-ohm, 1-watt
R_5 = 75,000-ohm, 1-watt
R_6 = 10,000-ohm potentiometer
R_7 = 30,000-ohm, $\frac{1}{2}$-watt
R_8 = 7,000-ohm, 10-watt
R_9 = 5,000-ohm, 10-watt

PLATE XI


NEON BULB
R_1
T_1
T_2
836
83S
T_3
L_1
L_2
C_1
C_2
C_3
C_4
C_5
+830 VOLTS
+350 VOLTS
VOLTMETER
VOLTMETER
R_2
R_3
R_4
R_5
R_6
R_7
R_8
R_9
2A3
2A3
6SJ7
X
Y
VR 90
VR 105
VR 150
+105 VOLTS
+150 VOLTS
115 VOLT 60 CYCLE A-C INPUT
T_4
T_5
5 VOLTS
6.3 VOLTS
6.3 VOLTS
700 VOLTS C.T.
FILAMENT DIAL LIGHT
24 VOLTS

such as a shorted quarter-wave section of transmission line would be a narrow band transformer and nullify the broad band characteristic of the antenna.

The third requirement was that the antenna have a nearly circular horizontal radiation pattern.

The fourth requirement was that the gain be as high as possible and still be consistent with the first three requirements.

Development

Since a wavelength is only slightly over a foot at 885 megacycles, a long antenna could be constructed without resulting in excessive physical size. A nonresonant antenna was especially desirable because it is broad band and also has high gain.

The biconical horn discussed by Barrow, Chu, and Jansen[1] fulfilled the requirements listed very satisfactorily. However, such an antenna constructed of sheet metal required some physical modification. Namely, the wind resistance had to be reduced, the antenna support simplified, and the roof eliminated.

As an approach to these problems an explanation for the evolution of the biconical antenna from a simpler type was searched for. Actually, the biconical horn is the horn resulting from the revolution of an ordinary flared rectangular horn about the apex of the pyramidal section. The biconical antenna may also be considered as a special shape of wave guide or as a

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[1] W. L. Barrow, L. J. Chu, and J. J. Jansen, "Biconical Electromagnetic Horns," _Proceedings of the I. R. E._, 27:769, December, 1939.

tapered coaxial line section. A discussion of the evolution of antennas of this type from tapered coaxial line sections has been given by Joseph M. Boyer.[1] However, none of these developments was applicable to the problem of simplifying the structure without materially altering the characteristics of the antenna. Consequently, a new approach was taken. It seemed that a biconical horn was essentially an infinite number of "V" antennae in parallel. If this were true, then an array of four "V" antennas in quadrature should produce a radiation pattern approximating that of a biconical horn.

A model antenna with elements 3 wavelengths long at 2400 megacycles was constructed and its radiation patterns plotted. The horizontal radiation pattern was a four-leafed rose. Next the number of elements per cone was increased to eight. This improved the horizontal radiation pattern, but there were still pronounced maximums off the end of each "V". However, the antenna wae beginning to behave like a biconical horn. In particular, it was noted that the spacing between the apexes of the cones affected the excitation. This effect was also observed by Barrow, Chu, and Jansen during experimentation with their biconical horn. The spacing influences the impedance of the exciting rod and thereby the energy transfer from the coxial line to the horn. In addition, the spacing influences directly the type of wave that may be excited. In general, higher order waves are easier to excite with greater separations of the cones. The vertical radiation pattern for the model antenna showed that a

[1] Joseph M. Boyer, "Discone--40 to 500 Mc Skywire," CQ, 5:11, July, 1949.

narrow beam was being radiated between the two cones, but considerable radiation in the vertical direction was also occurring.

The radiation along the vertical axis was approximately equal to the maximum horizontal radiation. The minor lobes from the individual "V" antennas were adding in phase to produce a major lobe. The primary difference between the model antenna composed of 8-element cones and a biconical horn was that the solid conducting plane of the cones in a biconical horn prevented all vertical radiation. As a result of this analysis, it was apparent that the biconical horn could be approximated by parallel "V's" only if sufficient number of them were used so that the cones appeared to be solid conducting sheets near their apexes.

The number of elements in the cones of the model antenna was increased to 16. Previous to this time the angle between the elements of each "V" had been 60 degrees, which is the optimum angle for a "V" antenna 3 wavelengths long. However, when the model antenna composed of 16-element cones was excited, it was necessary to decrease the angle to 40 degrees to bring the lobes together. The horizontal radiation pattern of this antenna was satisfactory but there was still appreciable vertical radiation. Another model with rod elements spaced radially every ten degrees was constructed. This model appeared to have the desired characteristics. To have a standard of comparison a biconical horn with the same dimensions was constructed of 1/2-inch mesh screen. The model antenna with 36 elements per cone compared very favorably with the biconical horn made of screen. The improvement resulting from a further increase in

the number of elements is small. Consequently it is possible to approximate a biconical horn made from conducting sheets by a biconical horn constructed of rod elements spaced every ten degrees radially. Construction in this manner reduces the wind resistance and eliminates the necessity of constructing a roof over the antenna.

Design

The design of the antenna followed the principles established by Barrow, Chu, and Jansen. The radiation pattern of a biconical horn is governed by the same parameters as the more common flared rectangular and circular horns. The gain is higher the longer the elements of the cone, and for each element length there is an optimum flare angle. The length of the cone elements was chosen as 3 feet for physical reasons. This gives an antenna that is approximately 6 feet in diameter and about as large as can be supported by a single pole. Elements of this length also cut economically from commercially available rod lengths of 12 feet. The cone elements are approximately 3 wavelengths long at 885 megacycles. For a length of 3 wavelengths the optimum flare angle is 40 degrees and the gain is approximately 15 decibels.

A biconical horn of this type may be excited with either TEM or $TE_{o,1}$ waves. TEM waves are fundamental to the horn and are the easier to excite. TEM waves produce vertical polarization while $TE_{o,1}$ produce horizontal polarization. $TE_{o,1}$ excitation may be produced by a number of dipole antennas bent into a

circle and disposed about the center in the equatorial plane. The balanced wire feed prevents the excitation of TM waves. TEM excitation was chosen and is to be accomplished by connecting the upper cone to the center conductor of the coaxial line feeding the antenna. The outer conductor will be connected to the lower cone. The spacing between the conical sections should be small to increase the attenuation of the higher order modes.

Physical Details

Aluminum was chosen for the antenna in preference to brass because it is cheaper and reduces the weight of the antenna. In the design the elements of the cones are 1/4-inch solid aluminum rods. They will screw into an aluminum hub 8 inches in diameter made from one-inch plate. The outer ends of the rods will also be threaded and locked by nuts into the 5/8-inch aluminum band that encloses the outer circumference of the conical section. The rods are spaced 10 degrees radially and are 6-1/4 inches apart at their outer ends. Both conical sections are identical. The two hubs are to be separated by a polystyrene block 3 inches in diameter. This block is 2 inches high and sets into a hole 1/2 inch deep milled into each hub. The polystyrene block serves as insulation between the two cones and also keeps the cones centered over each other. The center conductor of the coaxial line will extend through the center of this block. The upper conical section is to be attached to the lower section by paraffinated wooden strips connecting their outer rims.

EXPLANATION OF PLATE XII

Fig. 11. Assembled biconical horn. The paraffinated wooden braces between the outer rims of the conical sections are omitted for clarity.

Fig. 12. Antenna hub assembly.

Fig. 13. Antenna hub detail.

PLATE XII

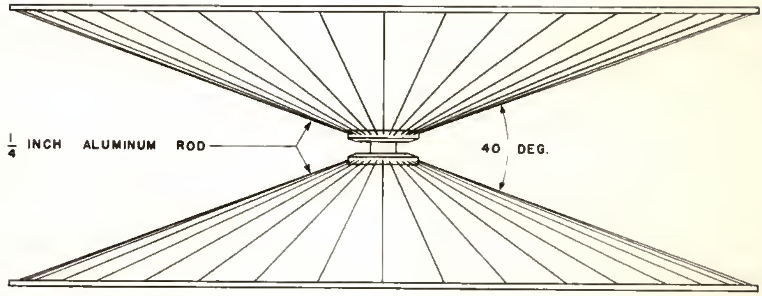

FIG. 11

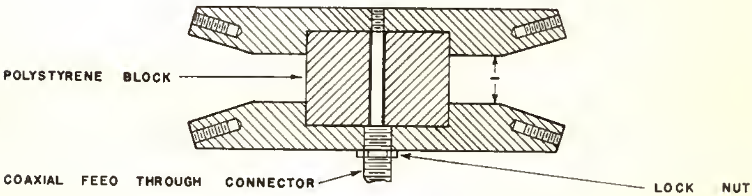

FIG. 12

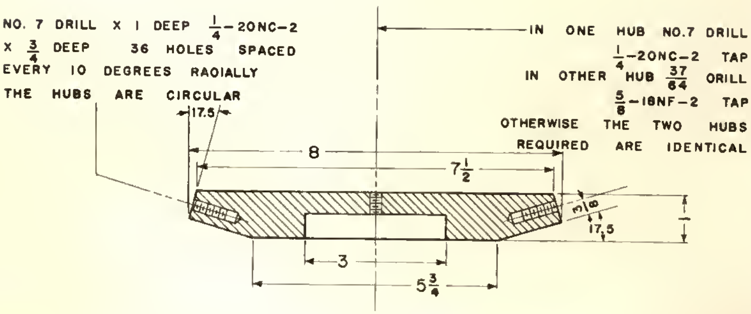

FIG. 13

CONCLUSIONS

During the preparation of this thesis it was noted that the effects of the unloaded shunt impedance of the plate tank circuit upon the operation of a very-high-frequency class C stage hae been neglected in most of the published literature. Instead the emphasis has been placed upon the Q of the tank circuits. This condition has probably developed because most authors have discussed lumped constant tank circuits in terms of their Q's. Once the convention was established, the tendency has been to continue to use this parameter in the discussion of ultra-high-frequency tank circuits. The Q of the tank coil and the unloaded shunt impedance of the tank circuit are directly related for lumped constant circuits. However, the bandwidth Q of a distributed constant tank circuit is not related to its shunt impedance.

The difference in the way these parameters vary may be shown by citing an example. The Q of an antiresonant section of transmission line is $(2\pi n f_o Z_o)/(Rc)$, and the input impedance of the section of transmission line ie $(8\ Z_o^2 f_o)/(nRc)$. Two important differencee in these equations can be noted. The input impedance ie directly proportional to the square of the characteristic impedance of the line, while the Q varies only as the first power of the characteristic impedance. Therefore a resonant line tank designed for maximum Q will not have as large a shunt impedance as a tank designed for maximum shunt impedance. The eecond important difference is the effect of making the resonant section of transmission line more than a

quarter wavelength long. If the resonant tank is a shorted three-quarter wavelsngth line, it will have a Q 3 times as large as that for a similar tank resonant at the same frequency but only a quarter wavelength long. However, the effect on the input impedance of the tank is the opposite. Consequently, designing a distributed constant tank circuit for maximum Q does not insure a high input impedance.

The band width Q of distributed constant circuits is generally high and sufficient selsctivity is obtained in the general case without special precaution. Unless the frequency stability of the circuit is sufficiently important to warrant some sacrifice in operating efficiency, distributed tank circuits should be designed with regard to their shunt impedance. This is the most useful characteristic of the resonator. It is the factor which tells the amount of input power that must be supplisd to the resonator to maintain a given voltage across the resonator input. Since the desired voltage across the resonator is determinsd by the class C design calculations for the stage, the shunt impedance of the resonator offers a method of evaluating the power loss in the resonator.

The effects of a resonator which has a known shunt impedance upon a class C stage may be calculated in two ways. The actual power loss may be determined from

$$\text{Average powsr loss} = \text{Voltage}^2/R_{\text{shunt}}$$

The voltage is the effective tank voltage from the class C design. This power loss may then be subtracted from the plate power output of the tube to give the net power output. Another

procedure for the evaluation of the effects of the resonator upon the class C stage is to calculate the tank circuit efficiency. The tank efficiency in terms of shunt impedances is

$$\frac{R_{shunt} - R_{loaded}}{R_{shunt}}$$

where R_{shunt} = unloaded shunt tank impedance

R_{loaded} = required tank impedance from class C design.

This equation shows the importance of a high unloaded shunt resistance of the tank circuit of a class C stage.

The tank efficiency may also be expressed by the general equation of output over input. The tank circuit input is the power output of the tube, or $(E_{bb} - E_{min})I_{ac}/2$, and the loaded tank impedance from class C design is $(E_{bb} - E_{min})/I_{ac}$. If the two equations for efficiency are equated and substitution made for the input and loaded tank impedance, the following equation results.

$$\text{Output} = \frac{(E_{bb} - E_{min})I_{ac}}{2} - \frac{(E_{bb} - E_{min})^2}{2\,R_{shunt}}$$

At very high frequencies the unloaded shunt resistance that can be obtained from lumped-constant circuits is fixed by the maximum ωLQ obtainable. The inductive reactance is limited by the interelectrode capacitance of the tube and the Q by physically realizable coils. For ultra high frequencies the unloaded shunt tank impedance is fixed once a given resonator is chosen.

I_{ac} and E_{min} were fixed during the class C design. As long as the peak cathode emission and the angle of plate current flow remain constant, which will occur if E_{min} remains constant and the necessary corrections are made in the grid bias, changes in

E_{bb} will not alter I_{ac}. Then E_{bb} is an independent variable in the equation for the tank output and may be changed without producing a change in the other terms on the right-hand side of the equation.

By differentiating tho output with respect to E_{bb} and setting the result equal to zero, a valuo of E_{bb} may be found which will produce maximum output for the given circuit. The result is

$$E_{bb} = E_{min} + \frac{R_{shunt} I_{ac}}{2}$$

If this result is substituted back into the original equations and the efficiency determined, it will be found to be 50 per cent, which is to be expected because this will be the condition for maximum power transfer.

At high frequencies R_{shunt} for a well designed tank circuit is so large that E_{bb} is limited by the tube rating before the above equality can be obtained. However, as the frequency increases, the obtainable shunt tank resistance from lumped-constant circuits decreases and it may become advisable to reduce E_{bb} in order to fulfill the equation.

For an example, consider a 6J6 dual-triode operating as a class C oscillator with the sections in push-pull. For a peak cathode emission of 75 milliamperes per section, E_{min} of 25 volts, and E_{max} of 8 volts, the peak value of the fundamental component of plate current is 25 milliamperes. First, typical operation of the oscillator at 150 megacycles with a plate-supply voltage of 250 volts is considered. $(E_{bb} - E_{min})/I_{ac}$ = 9,000 ohms, but this is tho requirod loaded shunt tank impedance for only one section. The total loaded shunt tank impedance required is

4 x 9,000 = 36,000 ohms for push-pull operation. Assuming an effective Q of 10 for the tank circuit, ωL is 3,600 ohms. A capacitance of 0.295 micromicrofarad is necessary to resonate this inductive reactance at 150 megacycles. However, the capacitance cannot be reduced to this value because the plate-to-cathode capacitances of the two sections in series is 0.25 micromicrofarad and in addition to this will be the stray wiring capacitance and the distributed capacitance of the coil appearing across the tank. For this example the total capacitance across the plate tank will be assumed to be 1.2 micromicrofarads. Then the capacitive reactance of the tank is 895 ohms at 150 megacycles. ωL is now 895 ohms and the effective Q of the tank must be increased to 40.2 to provide sufficient loaded tank impedance for class C design conditions. The unloaded Q of the tank circuit is approximately 125 at 150 megacycles. Therefore the unloaded shunt tank impedance is ωLQ = 112,000 ohms and the tank circuit efficiency 68 per cent. The power delivered to the tank circuit is $\frac{(E_{bb} - E_{min})I_{ac}}{2}$ = 2.69 watts per section. The total power input to the tank for push-pull is 5.38 watts and the power output 3.66 watts. After subtracting the driving power for the oscillator the net power output will be above 3 watts. The RCA tube manual gives 3.5 watts power output as typical power output for a 6J6. This portion of the example has illustrated the importance of considering the unloaded shunt impedance of the tank circuit.

The 6J6 oscillator frequency is now increased to 250 megacycles. Since the frequency does not enter into the class C de-

sign calculations, the peak value of the fundamental component of the plate current given in the preceding example will be unchanged. For a capacitance of 1.2 micromicrofarads across the tank the capacitive reactance at 250 megacycles is 532 ohms. Assuming an unloaded Q of 75 for the tank circuit, the unloaded shunt tank impedance is $\omega LQ = 40{,}000$ ohms. This tank impedance reduced for one section is 10,000 ohms.

$$E_{bb} = E_{min} + \frac{R_{shunt}\, I_{ac}}{2} = 150 \text{ volts}$$

Consequently at 250 megacycles, it is necessary to reduce the plate-supply voltage from the 250 volts used at 150 megacycles to 150 volts if the maximum output is to be obtained. The power delivered to the tank circuit is $(E_{bb}' - E_{min})I_{ac}/2 = 1.56$ watts per section, or 3.12 watts total. When E_{bb} fulfills the equality above the tank circuit, efficiency is 50 per cent; consequently the power output of the tank is 1.56 watts. For a plate-supply voltage of 150 volts the grid driving power is approximately 0.35 watt. Then the net power output of the oscillator at 250 megacycles is 1.2 watts. This exceeds slightly the rated output of 1 watt for a 6J6 push-pull oscillator at 250 megacycles. However, the example has illustrated the necessity of reducing the plate-supply voltage to obtain maximum output when using lumped-constant circuits at the higher frequencies.

In general, the unloaded shunt impedances obtainable by distributed constant circuits is sufficiently high that the tank efficiency is above 50 per cent and it is not necessary to reduce the plate voltage because of low tank impedance. Of course, the

more commonly given reasons for reduction in plate voltages still apply; i.e., transit time, etc. However, it may be necessary to operate frequency multiplier stages at reduced voltages because of the large loaded shunt impedance such stages require. Note should be taken of the fact that a tube operating single ended requires only one-fourth the loaded shunt tank impedance of two tubes operating under the same conditions in push-pull. Also a push-push doubler requires only one-half the shunt impedance that one section operating as a doubler would require. However, both sections operating in parallel as a doubler would require the same shunt impedance as the two sections operating push-push. In general, it may be said that tubes having large current capacities in comparison to their voltage ratings require lower loaded shunt tank impedances.

The fact that tho tank efficiency with distributed constants is usually above 50 per cent does not detract from the importance of designing the tank for high shunt impedanco. Designing shorted sections of transmission lines for maximum input impedance, i.e., establishing b/a ratios of 8.0 for parallel wire lines and 9.2 for coaxial lines,is not the only avenue open to the designer. The attenuation may also be reduced by increasing the size of the line conductors. It should be pointed out that maximum Q for sections of transmission lines occurs for a b/a ratio of 3.6 for coaxial lines and 4.0 for parallel wire linos. These are the conditions for minimum attenuation and are just one-half of the optimum ratios from the standpoint of input impedance. The input impedance of sections of transmission lines is approximately

proportional to the size of the conductors if the b/a ratio is held constant. Consequently, it is possible to adjust the input impedance of a shorted line by properly choosing the elements that compose the line section.

SUMMARY

The 885-megacycle television transmitter was designed to employ two 2C43 lighthouse tubes in push-pull in the final amplifier. The peak power output from design calculations is 18.3 watts. The tank circuits are shorted three-quarter wavelength parallel conductor transmission lines. Grounded grid construction is used. The final amplifier is to be plate modulated. Resistance coupling of the modulator to the final amplifier was used in the design to provide a modulator band pass of 5 megacycles.

A crystal-controlled oscillator at a frequency of 73.75 megacycles was incorporated into the exciter design. The second section of the 12AT7 dual-triode serves as a doubler to 147.5 megacycles. Lumped-constant tank circuits were used for the oscillator-doubler in this design. A 6J6 push-push doubler with a capacitively-loaded coaxial-line resonator increases the frequency to 295 megacycles. The frequency multiplication is completed by a 2C40 tripler. A 2C40 power amplifier at 885 megacycles completes the exciter. Both lighthouse stages were designed to use shorted flat-element transmission lines for tank circuits with grounded grid construction.

A biconical broadcast antenna is to be used with the transmitter. The conical sections of the antenna were designed to

be constructed of rod elements three wavelengths long and spaced ten degrees apart. Vertical radiation from the antenna is suppressed and the horizontal radiation pattern is circular. The antenna gain from design curves is 15 decibels. Because it is a nonresonant antenna it will have a broad band.

ACKNOWLEDGMENTS

The author is indebted to the members of the Department of Electrical Engineering of Kansas State College for their helpful encouragement and suggestions. In particular, the author wishes to thank Professor J. E. Wolfe for his invaluable assistance during the experimental phases of this project and for his criticisms of the manuscript.

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APPENDIX

PLATE-MODULATED FINAL AMPLIFIER

Class C Design

The following design was calculated in accordance with the procedure developed by Terman.[1] Two 2C43 lighthouse triodes are used in the final amplifier. The manufacturer's ratings for the 2C43 as given by the RCA Tube Manual are maximum plate dissipation 10 watts, maximum d-c plate voltage 450 volts, and maximum d-c plate current 36 milliamperes. The ARRL Handbook for 1949 lists 500 volts as maximum d-c plate voltage and 40 milliamperes as maximum d-c plate current. The RCA values were used since they are more conservative.

For oxide-coated cathodes the efficiency at normal operating temperatures is from 50 to 125 milliamperes per watt of heating power.[2] A cathode efficiency of 90 milliamperes per watt and a factor of safety of 2.5 were assumed for this design. The cathode heating power for the 2C43 is 5.68 watts and the peak cathode emission is 200 milliamperes.

From plate characteristics for the 2C43 a value of grid voltage and plate voltage was chosen to give 200 milliamperes total space current. For the point where E_{min} (plate) is 50 volts and E_{max} (grid) is 18 volts, the plate current is 120 milliamperes and the grid current is 80 milliamperes.

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[1] Frederick E. Terman, Radio Engineering, Third Edition (New York: McGraw-Hill), 1947, pp. 384-389.

[2] Frederick E. Terman, Radio Engineering, Second Edition (New York: McGraw-Hill), 1937, p.107.

The final amplifier was designed to operate good class C at the crest of the modulation cycle. The angle of plate current flow at crest conditions was chosen as 160 degrees. On the assumption that α equals 3/2, the d-c and fundamental frequency components of the total space current are 0.25 and 0.425, respectively, of the peak space current. The d-c component of the total space current is 50 milliamperes and the crest value of the fundamental frequency component of the total space current is 85 milliamperes. Using an amplification factor of 48 for the tube and a plate supply voltage of 450 volts, the grid bias required for an angle of plate current flow of 160 degrees is 13.4 volts. The angle of grid current flow is 130 degrees. The d-c and fundamental frequency components of the grid current are 0.18 and 0.33 times the peak value of the grid current, or 14.4 and 26.4 milliamperes, respectively. The d-c plate current is 50 - 14.4 = 35.6 milliamperes, and the fundamental frequency component is 85 - 26.4 = 58.6 milliamperes crest value. The power input to the plate is 16 watts. The power output is 11.7 watts and the plate dissipation is 4.3 watts. The loaded shunt tank impedance required is 6,820 ohms, and the grid driving power required is 0.415 watt.

The above design conditions are for one tube operating at the crest of the modulation cycle. The final amplifier operates push-pull, grounded grid. For two tubes the power output at the crest of the modulation cycle is 23.4 watts. The maximum d-c plate current is 71.2 milliamperes and the average d-c plate current is 35.6 milliamperes. The loaded shunt tank impedance re-

quired for design conditions is 4 x 6,820 = 27,280 ohms. The total driving power required is 3.5 watts including the power delivered to the cathode tank because of grounded grid construction.

Tank Circuit Design

The interelectrode capacitances of the 2C43 are as follows:

Grid-to-plate capacitance equals 1.7 micromicrofarads
Grid-to-cathode capacitance equals 2.8 micromicrofarads
Plate-to-cathode capacitance equals 0.02 micromicrofarad

Plate Tank. The plate line is 6 inches long to fit in the brass shield. However, the short is not an infinite conducting plane and contributes approximately an inch to the line length since the conductors are spaced 2 inches apart and the short is approximately the same size as the conductors. The plate caps of the tubes contribute another half inch to the line length, which makes the line effectively 7.5 inches or 0.19 meter long. At the frequency of 885 megacycles per second a wavelength is 0.339 meter. The plate line is 0.57 wavelength or 202 electrical degrees in length.

The plate line is constructed of 3/8-inch diameter tubing spaced 2 inches between centers. The characteristic impedance of the line is 272 ohms. The inductive reactance provided by a shorted transmission line more than one-half and less than three-fourths wavelength long is equal to the characteristic impedance of the line times the tangent of the line length in degrees.

$$X_L = Z_o \tan \theta = 272 \tan 202° = 109.5 \text{ ohms}$$

The capacitance required to resonate this inductive reactance at a frequency of 885 megacycles is 1.64 micromicrofarads. This capacitance is furnished by the grid-to-plate capacitance of the tubes and the tuning condenser. The grid-to-plate capacitances appear in series across the open end of the line for push-pull operation. This is 0.85 micromicrofarad of capacitance for two 2C43's. The remaining 0.8 micromicrofarad is contributed by tho tuning condenser.

Cathode Tank. Thc effective length of the cathode line is approximately 7 inches or 0.178 meter. At the frequency of 885 megacycles the cathode line is 0.525 wavelength or 189 electrical degrees in length.

The cathode line is constructed of 1 1/2-inch diameter tubing spaced 2 inches between centers. The characteristic impedance of the line is 61 ohms.

$$X_L = Z_o \tan \theta = 61 \tan 189° = 9.65 \text{ ohms}$$

The capacitance required to resonate this inductive reactance at 885 megacycles is 18.6 micromicrofarads. Of this capacitance 1.4 micromicrofarads arc contributed by thc grid-to-cathode capacitances of the tubes in scries and the remaining 17.2 micromicrofarads are furnished by the tuning condenser.

Plate Tank Efficiency

The efficiency of the plate tank circuit of a class C stage has been defined as $(Q_1 - Q_{eff})/Q_1$, where Q_1 is the figure of merit of tho unloaded tank and Q_{eff} is the effective Q of the

loaded tank circuit. This equation is accurate only for lumped-constant circuits where the shunt impedances of the tank are directly related to the Q of the tank circuit. This equation may be written in a more general form which is applicable to tanks composed of either lumped constants or distributed constants.

$$\text{Tank efficiency} = \frac{(R_{shunt} - R_{loaded})}{R_{shunt}}$$

where R_{shunt} = the shunt impedance of the unloaded tank circuit

R_{loaded} = the value of plate load impedance for class C design conditions.

Consequently the efficiency of the plate tank circuit is dependent upon the attainable unloaded shunt impedance of the tank.

The input impedance to a shorted transmission line that is an odd multiple of quarter wavelengths long is a large pure resistance.[1]

$$R_{sc} = \frac{Z_o}{\tanh \alpha \ell} = \frac{Z_o}{\alpha \ell} \text{ for small values of } \alpha \ell .$$

where Z_o = characteristic impedance of the line
α = attenuation constant in nepers per meter
ℓ = line length in meters = $cn/4f$
c = velocity of light in meters per second
n = number of quarter-wave sections in the line
f = frequency at which the line is antiresonant.

The maximum shunt impedance that a tank composed of a shorted section of transmission line an odd multiple of quarter wavelengths long can have is $4\, Z_o f/\alpha cn$. However, in the actual case the lines which form the external tank circuits are not an

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[1] Robert I. Sarbacher and William A. Edson, Hyper and Ultrahigh Frequency Engineering (New York: John Wiley & Sons), 1947, p. 349.

odd multiple of quarter wavelengths long but are shortened by the capacitive loading of the tubes. The reduction in shunt impedance of capacitively-loaded shorted transmission lines may be taken into account by applying a correction factor to the shunt impedance that the tank would have if it were an integral number of quarter wavelengths long.

The shunt resistance of an ordinary resonant circuit may be defined as[1]

$$R_{shunt} = \frac{\text{voltage}^2}{2 \text{ x energy lost per second}}$$

This definition may also be applied to other types of resonators.

When a shorted section of a transmission line is loaded with a capacitance across the open end to produce a resonant tank, standing waves will exist on the line and the magnitude of the voltage and current at any point along the line is given approximately by the following equations.

$$E' = E_{max} \sin (2\pi x)/\lambda$$

$$I' = I_{max} \cos (2\pi x)/\lambda$$

where E_{max} = the voltage a quarter wavelength from the short
I_{max} = the current at the short
x = the distance from the point under consideration to the shorted end in wavelengths.

The energy loss per second in an infinitesimal length, dx, is $I'^2R\,dx$ and the energy loss per second for a line x wavelengths long is $\int_0^x I'^2R\,dx$, where R is the resistance of the line per wavelength.

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[1] Theodore Moreno, Microwave Transmission Design Data (New York: McGraw-Hill), 1948, p. 215.

$$\text{Energy loss/sec.} = I_{max}^2 \frac{R\lambda}{2\pi}\left(\frac{\pi x}{\lambda} + \tfrac{1}{2}\sin\frac{2\pi x}{\lambda}\cos\frac{2\pi x}{\lambda}\right)$$

$$R_{shunt} = \frac{\pi E_{max}^2 \sin\frac{2\pi x}{\lambda}}{R\lambda I_{max}^2\left(\frac{\pi x}{\lambda}\csc\frac{2\pi x}{\lambda} + \tfrac{1}{2}\cos\frac{2\pi x}{\lambda}\right)}$$

When the line is a quarter wavelength long

$$R_{shunt} = R_{sc\frac{\lambda}{4}} = \frac{4 E_{max}^2}{R\lambda I_{max}^2}$$

For a line any length less than a quarter wave long

$$R_{shunt} = R_{sc\frac{\lambda}{4}} \frac{\pi}{2} \frac{\sin\frac{2\pi x}{\lambda}}{\frac{2\pi x}{\lambda}\operatorname{cosec}\frac{2\pi x}{\lambda} + \cos\frac{2\pi x}{\lambda}}$$

When the line is three-fourths of a wavelength long

$$R_{shunt} = R_{sc\frac{\lambda}{4}} = \frac{4 E_{max}^2}{3 R\lambda I_{max}^2} = \frac{R_{sc\frac{\lambda}{4}}}{3}$$

This agrees with the fact established previously that the shunt impedance of a line an odd multiple of quarter wavelengths long varies inversely with the number of quarter-wave sections in the line. For lines between a half and three-quarters of a wavelength long

$$R_{shunt} = R_{sc\frac{3\lambda}{4}} \frac{3\pi}{2} \frac{\sin\frac{2\pi x}{\lambda}}{\frac{2\pi x}{\lambda}\operatorname{cosec}\frac{2\pi x}{\lambda} + \cos\frac{2\pi x}{\lambda}}$$

The ratio of the shunt impedance to what it would be if the line were a quarter or three-quarters of a wavelength long have been

calculated and are plotted versus the line length in wavelengths in Plate XIII. The results of these calculations agree very closely with those obtained by Nergaard and Salzberg.[1] Their results were arrived at by different method and were substantiated by experimental evidence.

The shunt impedance that the plate tank of the final amplifier would have if the line were three-quarters of a wavelength long was computed with the equation

$$R_{sc} = \frac{4\ Z_o f}{\alpha\ cn}$$

For transmission lines where the conductance per unit length may be assumed to be zero, the attenuation constant is $R/2\ Z_o$, where R is the loop resistance of the line per unit length. For parallel wire lines separated by air

$$\alpha = \frac{\frac{1}{\pi a}\sqrt{\frac{\pi f \mu}{\sigma}}}{2 \times 120 \log_e \frac{b-a}{a}} \quad \text{nepers per meter}$$

where a = radius of each conductor in meters
b = distance between the centers of the conductors in meters
f = frequency in cycles per second
μ = permeability of the line material, approximately equal to the permeability of free space for most conductors
σ = conductivity of the line material in mhos per meter.

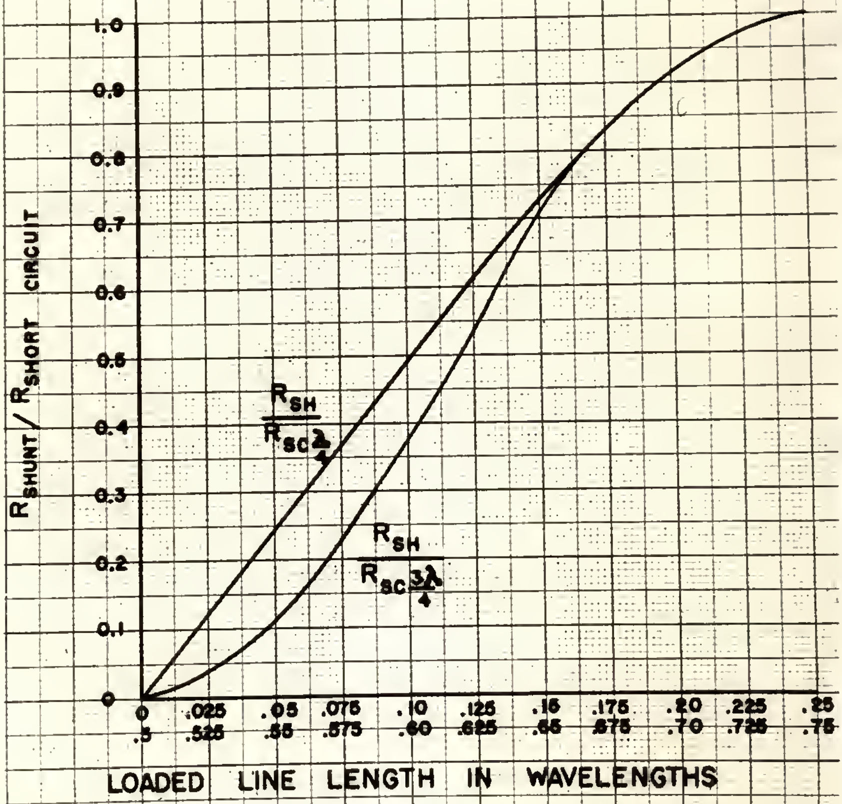
For silver-plated lines 3/8 inch in diameter, spaced 2 inches apart, and three-quarters wavelengths long

[1] L. S. Nergaard and Bernard Salzberg, "Resonant Impedance of Transmission Lines," Proceedings of the I. R. E., 22:579, September, 1939.

EXPLANATION OF PLATE XIII

Plot of the ratio of the shunt resistance of a resonant shorted section of transmission line shortened by capacitive loading, R_{shunt}, to the shunt resistance of a shorted section of the same transmission line an odd number of quarter wavelengths long, $R_{short\ circuit}$, versus the loaded line length.

PLATE XIII



1.0
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
R SHUNT / R SHORT CIRCUIT
R SH / R SC λ/4
R SH / R SC 3λ/4
0 .025 .05 .075 .10 .125 .15 .175 .20 .225 .25
.5 .525 .55 .575 .60 .625 .65 .675 .70 .725 .75
LOADED LINE LENGTH IN WAVELENGTHS

$a = 4.76 \times 10^{-3}$ meters
$b = 5.08 \times 10^{-2}$ meters
$f = 885 \times 10^{6}$ cycles per second
$\mu = 1.257 \times 10^{-6}$ henries per meter
$\sigma = 6.1 \times 10^{7}$ mhos per meter
$\alpha = 9.39 \times 10^{-4}$ nepers per meter
$R_{sc\frac{3\lambda}{4}} = 1.14 \times 10^{6}$ ohms

The shunt resistance of the final amplifier plate tank which is 0.57 wavelength long is 0.11 $R_{sc\frac{3\lambda}{4}}$ from Plate XIII, or 1.25×10^{5} ohms. The plate-tank efficiency of the final amplifier is 78.4 per cent and the power output 18.3 watts.

RESISTANCE-COUPLED MODULATOR

Resistance Modulation Design Procedure

For resistance coupling of the modulator to the class C stage the modulator tube operates in parallel with the class C stage. Also the class C stage is adjusted for a linear characteristic of d-c current versus d-c voltage over the range of operation. Therefore, the class C stage appears to the modulator as a resistance equal to E_b/I_b. The basic modulator circuit given in Plate IV shows this simplification.

Since it is not feasible to reduce the plate voltage of the modulator tube to zero, the cathodes of the final amplifier tubes are operated at a constant potential above ground. This is accomplished by placing a dropping resistor, R_d, between the cathode of the final amplifier and ground. The output of a regulated power supply maintains a constant current in R_d. Therefore the plate-to-cathode voltage of the final amplifier is always less than the plate-to-cathode voltage of the modulator. From the

characteristics of the proposed modulator tube this voltage below which it is not feasible to reduce the plate voltage is chosen and will be called E_{min}. Currents and voltages for the class C amplifier will be designated by the subscript A. Then to allow E_A to be reduced to zero for 100 per cent modulation, the voltage drop in R_d, E_d, must equal E_{min}.

Since the modulator tube is in parallel with the class C stage, E_{max} is fixed as $E_{min} + E_{A\,max}$. Now writing the loop equation for the left loop of the simplified modulator circuit,

$$E_{bb} - R_L(I_A + I) - E_A - E_d = 0$$

but $\qquad E_d = E_{min} \qquad$ and $\qquad E_A = I_A R_A$

Then $\qquad (E_{bb} - E_{min}) - I_A(R_L + R_A) - IR_L = 0$

This is a general equation which must be true at every instant. Now take the maximum and minimum conditions

$$(E_{bb} - E_{min}) - I_{A\,max}(R_L + R_A) - I_{min}R_L = 0$$

$$(E_{bb} - E_{min}) - I_{A\,min}(R_L + R_A) - I_{max}R_L = 0$$

But $I_{A\,min}$ is equal to zoro for 100 per cent modulation. Then by equating the two equations and solving for R_L,

$$R_L = \frac{E_{A\,max}}{(I_{max} - I_{min} - I_{A\,max})}$$

In this equation $I_{A\,max}$ and $E_{A\,max}$ are fixed by the class C design.

The values of I_{max} and I_{min} are chosen in accordance with the same principles that apply to class A power amplifiers. The only difference is that the plate voltage swing has been fixed by the previously determined values of E_{max} and E_{min}. After the values of I_{max} and I_{min} which give an operating point where the

plate dissipation is not excessive has been chosen, the value of R_L may be determined.

From the equation for maximum current in the modulator the required plate-supply voltage may be evaluated.

$$E_{bb} = E_{min} + I_{max} R_L$$

Modulator Design

This design follows the design procedure outlined and uses a 715A (W.E. 5D21) tetrode as the modulator tube. From the plate characteristics which are shown in Plate XIV, E_{min} was chosen as 100 volts. Then $E_{max} = E_{A\ max} + E_{min} = 550$ volts. I_{max} and I_{min} were chosen as 325 and 50 milliamperes, respectively. These conditions occur for an operating point at

$$\begin{aligned} E_{c1} &= -17.5 \text{ volts} \\ E_{c2} &= 150 \text{ volts} \\ E_{bo} &= 330 \text{ volts} \\ I_{bo} &= 180 \text{ ma} \\ E_g &= 35 \text{ volts peak-to-peak} \end{aligned}$$

Substituting these values in the equations developed previously gives

$$R_L = 2{,}200 \text{ ohms} \quad \text{and} \quad E_{bb} = 830 \text{ volts}$$

The a-c load resistance on the modulator is

$$\frac{(E_{max} - E_{min})}{(I_{max} - I_{min})} = 1{,}640 \text{ ohms}$$

Modulator Band Pass

The limiting factor to the high-frequency response of the modulator is the capacitance from plate to ground appearing across the output since the a-c load resistance has been fixed

EXPLANATION OF PLATE XIV

Plate characteristic of modulator showing a-c load line.

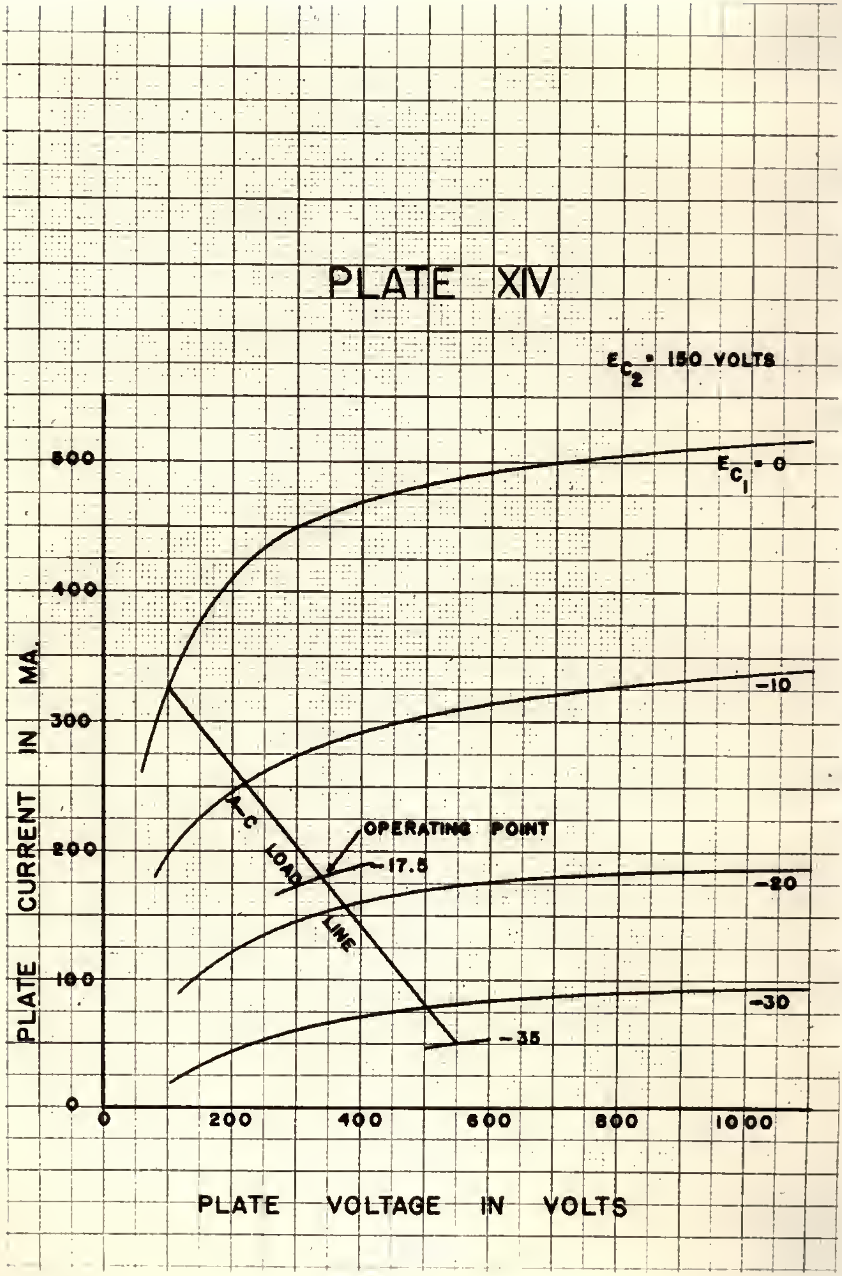


PLATE XIV
E_{C_2} = 150 VOLTS
E_{C_1} = 0
-10
-20
-30
-35
-17.5
OPERATING POINT
A-C LOAD LINE
PLATE CURRENT IN MA.
500
400
300
200
100
0
0
200
400
600
800
1000
PLATE VOLTAGE IN VOLTS

by the modulator design. This capacitance is the sum of the plate-to-cathode capacitance of the modulator tube, the stray capacitances due to the wiring, the capacitance of the plate tank of the final amplifier to ground, and the grid-to-plate capacitance of final amplifier tubes.

C_{pk} for 715A is 8 micromicrofarads
C_{gp} for one 2C43 is 1.7 micromicrofarads
$C_{stray\ wiring}$ assumed to be 8 micromicrofarads

The total capacitance across the output of the modulator is approximately 19.4 micromicrofarads. The a-c load resistance from the modulator design is 1,640 ohms.

$f_o = 1/2\pi RC = 5.0$ megacycles

$L = CR^2/2 = 26$ microhenries of shunt peaking inductance

Equations from Principles of Television Engineering.[1]

Harmonic Distortion

From the plate characteristic the following points were taken for making a five-point harmonic analysis of the output.[2]

$I_{max} = 325$ ma
$I_{+\frac{1}{2}} = 260$
$I_b = 180$
$I_{-\frac{1}{2}} = 105$
$I_{min} = 50$

By the procedure given in the reference cited, the following results were obtained.

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[1] Donald G. Fink, Principles of Television Engineering (New York: McGraw-Hill), 1940, p. 222.

[2] Jacob Millman and Samuel Seely, Electronics (New York: McGraw-Hill), 1941, p. 543.

D-c component of output wave = 4 milliamperes
Fundamental component of output wave = 143 ma
Second-harmonic component = 4 ma
Third-harmonic component = -6 ma
Fourth-harmonic component = -0.5 ma

The total distortion is approximately 5 per cent. The plate dissipation with sinusoidal signal is 60.7 watts.

Alternative Modulator Design Procedure

The equivalent circuit shown in Plate IV may be used to obtain an equation for the load resistance required. The a-c load resistance on the modulator is the parallel combination of the resistance presented by the class C stage and the load resistance. The resistance presented to the modulator by the class C stage is $\frac{E_{A\ max}}{I_{A\ max}}$. The a-c load resistance on the modulator is also $\frac{(E_{max} - E_{min})}{(I_{max} - I_{min})}$, which is equal to $\frac{E_{A\ max}}{(I_{max} - I_{min})}$.

Then

$$\frac{E_{A\ max}}{(I_{max} - I_{min})} = \frac{R_L\left(\frac{E_{A\ max}}{I_{A\ max}}\right)}{\left[R_L + \left(\frac{E_{A\ max}}{I_{A\ max}}\right)\right]}$$

Solving this equation for R_L

$$R_L = \frac{E_{A\ max}}{(I_{max} - I_{min} - I_{A\ max})}$$

This is the same result as was obtained in the preceding development.

The required plate-supply voltage may be obtained by use of the d-c load line for the modulator. The intercept of the d-c load line on the current axis of the plate characteristics is the

value of current flowing through the tube if its plate-to-cathode voltage were reducod to zero. This intercept is

$$\frac{(E_{bb}R_A + E_dR_L)}{R_AR_L}.$$

The intercept of the d-c load line on the voltage axis occurs whon no current passes through tho tube. The voltage intercept is

$$E_{bb} - \frac{R_L(E_{bb} - E_d)}{(R_A + R_L)}$$

From these two intercepts the slope of the d-c load line and the d-c load resistance can be determined. The d-c load resistance on the modulator is $R_AR_L/R_A + R_L)$, which is the same as the a-c load resistance. This is to be expected since the load on the modulator is composed of resistances. Consequently the modulator operates as a conventional class A power amplifier with a resistance load of $R_AR_L/(R_A + R_L)$ ohms and a power supply voltage of

$$\frac{(E_{bb}R_A + E_dR_L)}{(R_A + R_L)} \text{ volts.}$$

The voltage at the operating point is the voltage axis intercept of the d-c load line minus the voltage drop in the d-c load resistance due to the plate current flowing at the operating point.

$$E_{bo} = \frac{(E_{bb}R_A + E_dR_L)}{(R_A + R_L)} - \frac{I_{bo}R_AR_L}{(R_A + R_L)}.$$

When this equation is solved for the plate-supply voltage, E_{bb}, the following equation results

$$E_{bb} = E_{bo} + I_{bo}R_L + \frac{(E_{bo} - E_d)R_L}{R_A}$$

Substituting in the values for the operating point chosen gives the required plate supply voltage as 830 volts which is the same result as obtained by the other modulator design procedure.

73.75-MEGACYCLE OSCILLATOR

Class C Design

The following oscillator design was calculated in accordance with the procedure developed by Terman.[1] One-half of a 12AT7 dual triode is used as the oscillator tube. The manufacturer's ratings for one section of a 12AT7 are maximum plate dissipation 2.5 watts and maximum d-c plate voltage 300 volts.

This design is based upon a peak-cathode emission of 40 milliamperes and a plate supply voltage of 105 volts. The angle of plate current flow was chosen as 140 degrees. From plate characteristics for the 12AT7 a value of grid voltage and plate voltage was chosen to give 40 milliamperes total space current. For the point where E_{min} (plate) is 25 volts and E_{max} (grid) is 5 volts, the plate current is 25 milliamperes and the grid current is 15 milliamperes.

On the assumption that $\propto$ equals 3/2, the d-c and fundamental frequency components of the total space current are 0.225 and 0.39, respectively, of the peak space current. The d-c com-

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[1] Frederick E. Terman, Radio Engineering, Third Edition (New York: McGraw-Hill), 1947, p. 412.

ponent of the total space current is 9 milliamperes and the crest value of the fundamental frequency component of the total space current is 15.6 milliamperes. Using an amplification factor of 35 for the tube, the grid bias required for an angle of plate current flow of 140 degrees with a plate voltage of 105 volts is 5.97 volts. The angle of grid current flow is 114 degrees. The d-c and fundamental frequency components of the grid current are 0.17 and 0.31 times the peak value of grid current, or 2.55 and 4.65 milliamperes, respectively. The d-c plate current is 9 - 2.55 = 6.45 milliamperes, and the fundamental frequency component of plate current is 15.6 - 4.65 = 11 milliamperes crest value. The power input to the plate is 0.677 watt. The power developed in the plate tank is 0.440 watt and the plate dissipation is 0.237 watt. The grid driving power required is 0.0255 watt, which gives a net power output of the oscillator, neglecting tank losses, of 0.414 watt. The grid-leak resistance required is 2,340 ohms. The loaded plate tank impedance necessary to give these design conditions is 7,200 ohms.

Tank Circuit Design

For good plate tank efficiency the loaded Q of the plate tank should be small in comparison to the unloaded Q. However, for sharp resonance the loaded Q should not be less than 10. Assuming an effective Q of 10,

$$Z_p = \omega L\, Q_{eff} = 7{,}200 = 10\ \omega L$$

The capacitance required to resonate this inductive reactance at a frequency of 73.75 megacycles is 3 micromicrofarads. Since the

second section of the 12AT7 was capacitively coupled, both the output capacitance of the first section and the input capacitance of the second section appear across the plate tank. The output capacitance of the first section is 0.45 micromicrofarad and the input capacitance of the second section is 2.5 micromicrofarads. This is sufficient tuning capacitance and an inductance resonating the tube capacitances would give the best tank efficiency. However, to make adjustment of the oscillator easier and especially to simplify checking for crystal-controlled oscillation, a plate tank tuning condenser was included. This condenser was a 3-17-micromicrofarad midget variable. Assuming a setting of this tuning condenser of 10 micromicrofarads for design conditions, the capacitance across the plate tank becomes 15 micromicrofarads. In order to resonate this capacitance the inductive reactance of the tank coil must be smaller than 720 ohms. In fact, the inductive reactance of the tank coil must be 144 ohms to resonate 15 micromicrofarads at 73.75 megacycles. Now the effective Q of the tank circuit cannot be reduced to 10 and still provide sufficient plate tank impedance. The plate tank operates with an effective Q of 50. Assuming an unloaded Q of 150 for the tank coil, the plate tank efficiency is

$$(Q - Q_{eff})/Q = 66.7 \text{ per cent.}$$

The oscillator power output = 0.667 x 0.414 = 0.276 watt. The inductance of the plate tank coil is 0.31 microhenry.

12AT7 DOUBLER FROM 73.75 MC TO 147.5 MC

Class C Design

The following harmonic generator design follows the design procedure outlined by Terman.[1] The tube ratings are the same as those listed for the 12AT7 oscillator. This design uses a peak-cathode emission of 65 milliamperes, a plate supply voltage of 105 volts, and an angle of plate current flow of 100 degrees.

From the plate characteristics for the 12AT7 a value of grid voltage and plate voltage was chosen to give 65 milliamperes total space current. For the point where E_{min} (plate) is 25 volts and E_{max} (grid) is 8 volts, the plate current is 40 milliamperes and the grid current is 25 milliamperes.

On the assumption that α equals 3/2, the d-c and second-harmonic component of the total space current are 0.165 and 0.25, respectively, of the peak space current. The d-c component of the total space current is 10.4 milliamperes and the crest value of the second-harmonic component of the total space current is 16.2 ma. Using an amplification factor of 35 for the tube, the grid bias required for a doubler operating with a plate voltage of 105 volts and an angle of plate current flow of 100 degrees is 20.6 volts. The angle of grid current flow is 88 degrees. The d-c component of the grid current is 0.13 times the peak value of the grid current, or 3.25 milliamperes. The fundamental

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[1] Frederick E. Terman, Radio Engineering, Third Edition (New York: McGraw-Hill), 1947, p. 394.

frequency component of the grid current is 0.245 x 25 = 6.13 milliamperes. The second-harmonic component of the grid current is 0.215 x 25 = 5.38 milliamperes crest value. The d-c plate current 10.4 - 3.25 = 7.15 milliamperes and the second-harmonic component of the plate current is 16.2 - 5.38 = 10.8 milliamperes crest value.

The power input to the plate is 0.75 watt. The second-harmonic power developed in the plate tank is 0.432 watt and the plate dissipation is 0.318 watt. The grid driving power required is 0.088 watt. The grid-leak resistance is 6,550 ohms. The loaded tank impedance necessary to give these design conditions is 7,400 ohms.

Tank Circuit Design

Since the coupling loop to the next stage is tuned to series resonance and does not reflect reactance into the doubler plate tank, the capacitance across the tank is the output capacitance of the tube plus the stray-wiring capacitance and the capacitance of the tank-tuning condenser. A condenser was used to tune the tank even though it results in reduced tank efficiency. The tuning condenser is a Cardwell Trim-Air Midget type ZR-10AS, 1.2-10 micromicrofarads variable. For a total capacitance of 5 micromicrofarads across the plate tank the inductive reactance required for resonance at 147.5 megacycles is 216 ohms. This required a loaded Q of 34.2 for the tank circuit to provide sufficient tank impedance for the design conditions. Assuming an unloaded Q of 125 for the plate tank coil, the plate tank effi-

ciency is 72.9 per cent. The doubler power output is 0.729 x 0.432 = 0.314 watt. The inductance of the plate tank coil is 0.233 microhenry.

6J6 PUSH-PUSH DOUBLER FROM 147.5 MC TO 295 MC

Class C Design

This harmonic generator design also follows the design procedure developed by Terman. The design is first made for one section of the dual-triode. The manufacturer's ratings for one section of the 6J6 are maximum plate dissipation 1.5 watts, maximum d-c plate current 15 milliamperes, and maximum d-c plate voltage 300 volts. A peak-cathode emission of 64 milliamperes was used in this design. The angle of plate current flow was chosen as 120 degrees and the plate supply voltage was 150 volts.

From the plate characteristics for the 6J6 a value of grid voltage and plate voltage was chosen to give 64 milliamperes total space current. For the point where E_{min} (plate) is 25 volts and E_{max} (grid) is 7 volts, the plate current is 40 milliamperes and the grid current is 24 milliamperes.

On the assumption that α equals 3/2, the d-c and second-harmonic component of the total space current are 0.19 and 0.265, respectively, of the peak space current. The d-c component of the total space current is 12.15 milliamperes and the crest value of the second-harmonic component of the total space current is 17 milliamperes. Using an amplification factor of 38 for the tube, the grid bias required for a doubler operating with a plate voltage of 150 volts and an angle of plate current flow of 120

degrees is 11.6 volts. The angle of grid current flow is 107 degrees. Tho d-c component of the grid current is 0.155 times the peak value of the grid current, or 3.72 milliamperes. The fundamental frequency component of the grid current is 0.29 x 24 = 6.95 milliamperes. The second-harmonic component of the grid current is 0.24 x 24 = 5.76 milliamperes crest value. The d-c plate current is 12.15 - 3.72 = 8.43 milliamperes and the second-harmonic component of the plate current is 17 - 5.76 = 11.24 milliamperes crest value.

The power input to the plate is 1.265 watts. The second-harmonio power developed in the plate tank is 0.703 watt and the plate dissipation is 0.562 watt. The grid driving power required is 0.0681 watt and the loaded tank impedance necessary for the design conditions is 11,100 ohms. These values are for one section of tho 6J6 only.

For both sections of the 6J6 operating in a push-push doubler for the second harmonic power output is 1.4 watts. The grid driving power required for both sections is 0.136 watt and the grid-leak resistance is 1,560 ohms. Since the plates of the two sections operate in parallel the loaded tank impedance required for design conditions is only 5,550 ohms.

Tank Circuit Design

The tank circuit of the 6J6 push-push doubler is composed of a capacitively-loaded coaxial-line resonator. This resonator is type D-223920 manufactured by the Lapp Insulator Company. The high-impedance input tap on the center conductor is brought out

3 1/8 inches from the shorted end. The output tap on the center conductor is 7/8 inch from the short. The total length of the center conductor is 3 1/2 inches and the loading capacitance is adjustable for tuning by screwing a cap on the outer cylinder of the resonator. The length of the center conductor is considerably less than a quarter wave length at 295 megacycles; therefore the loading capacitance required for tuning the resonator will not be negligible. When appreciable loading capacitance is introduced at the end of the line, the resonant frequency of a capacitively-loaded coaxial-line resonator is approximately the resonant frequency of the loading capacitance and the inductance of the shorted coaxial transmission line which composes the remainder of the resonator. The center conductor of the resonator is 0.0876 wavelength long or 31.5 degrees in length. The characteristic impedance of the coaxial line portion of the resonator is 83.4 ohms. The inductive reactance provided by this shorted section of coaxial transmission line is $Z_0 \tan \theta = 51.1$ ohms. The loading capacitanco required to resonate this inductive reactance at a frequency of 295 megacycles is 10.56 micromicrofarads. The tube interelectrode capacitances contribute 1 micromicrofarad of this capacitance because the two plates are in parallel. The remaining 9.56 micromicrofarads of capacitance must be provided by the loading capacitance at the end of the resonator. For a loading condenser plate 7/8 inch in diameter, the gap required between the condenser plate and the cap on the outer cylinder to provide the necessary capacitance is 0.0141 inch.

Plate Tank Efficiency

The tank efficiency was calculated by the same method that was used for the final amplifier plate tank. The input resistance of a shorted section of transmission line a quarter wave long is 4 $Z_o f/\alpha c$. The attenuation constant, α , for a coaxial transmission line with finite conductivity is[1]

$$\frac{1}{2}\sqrt{\frac{\pi f \mu_2 \varepsilon_1}{\sigma_2 \mu_1}}\left(\frac{1}{a}+\frac{1}{b}\right)\frac{1}{\log_e \frac{b}{a}}$$

where a = radius of the center conductor in meters
b = inner radius of the outer cylinder
ε_1 = dielectric constant of insulating medium
μ_1 = permeability of the insulating medium
μ_2 = permeability of the conductors
σ_2 = conductivity of the conductors

For the silver-plated capacitively-loaded coaxial-line resonator used for the tank

a = 0.0032 meter
b = 0.0127 meter
$\varepsilon_1 = 8.854 \times 10^{-12}$ farads per meter
$\mu_1 = \mu_2 = 1.257 \times 10^{-6}$ henry per meter
$\sigma_2 = 6.1 \times 10^{7}$ mhos per meter
$f = 295 \times 10^{6}$ cycles per second
$\alpha = 1.64 \times 10^{-3}$ nepers per meter

The characteristic impedance of a coaxial line is $(377/2\pi)\log_e(b/a)$ which is 83.4 ohms for the line used. Then the shunt impedance of a quarter wave line is

$$R_{sc\frac{\lambda}{4}} = 4\, Z_o f/\alpha c = 2 \times 10^{5} \text{ ohms}$$

However, the line is not a quarter wave long and this reduction

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[1] Robert I. Sarbacher and William A. Edson, Hyper and Ultra-high Frequency Engineering (New York; John Wiley & Sons), 1947, p. 284.

in length reduces the shunt impedance as previously calculated. From Plate XIII, the shunt resistance of a line 0.0876 wavelength long is 0.425 of what the shunt resistance would be if it were a quarter wavelength. The maximum shunt resistance of the resonator used for the 6J6 push-push doubler tank is $0.425 \times 2 \times 10^5 = 8.5 \times 10^4$ ohms; but the tube is tapped down on the line which still further reduces the unloaded shunt resistance of the resonator.

The effect of tapping down on a shorted transmission line on which standing waves exist is to transform the impedance by a nearly perfect transformer. For a perfect transformer, $V_1n_2 = V_2n_1$ and $I_1n_1 = I_2n_2$. By dividing the first equation by the second, the ratio of the input to output impedance may be determined. $Z_{in}/Z_{out} = (n_1/n_2)^2 = (V_1/V_2)^2$. For a shorted line with standing waves, the voltage is approximately $V_{max} \sin \theta$ where θ is measured from the short. If the impedance that the tube sees or, in other words, the impedance at the tap, is designated by Z_{in}, and the impedance at the open end of the line is called Z_{out}, then

$$Z_{in} = Z_{out} \left(\frac{\sin^2 \theta_1}{\sin^2 \theta_2} \right)$$

where θ_1 = distance from the short to the tap in degrees

θ_2 = length of the line in electrical degrees

The input impedance for the resonator at the open end, Z_{out}, has been shown to be 8.5×10^4 ohms, and θ_2 is 31.5 degrees. The input tap to the resonator is located 3 1/8 inches from the shorted end which is 28.1 degrees. The input resistance of the resonator at the high-impedance tap is 6.94×10^4 ohms.

The efficiency of the plate tank is

$$\frac{(R_{shunt} - R_{loaded})}{(R_{shunt})} .$$

The loaded tank impedance required for the class C design conditions is 5.55 x 10^3 ohms. The plate tank efficiency is 92 per cent and the power output of the doubler is 1.29 watts.

2C40 TRIPLER FROM 295 MC TO 885 MC

Class C Design

The manufacturer's ratings for the 2C40 lighthouse triode are maximum plate dissipation 5 watts, maximum d-c plate current 22 milliamperes, and maximum d-c plate voltage 450 volts. A peak-cathode emission of 116 milliamperes was used in this design. The angle of plate current flow was chosen as 90 degrees to increase the third-harmonic component of the space current. The plate supply voltage used was 350 volts.

From the plate characteristics for the 2C40 a value of grid voltage and plate voltage was chosen to give 116 milliamperes total space current. For the point where E_{min} (plate) is 50 volts and E_{max} (grid) is 12 volts, the plate current is 63 milliamperes and the grid current is 53 milliamperes.

On the assumption that α equals 3/2, the d-c component of the total space current is 0.145 x 116 = 16.8 milliamperes. The crest value of the fundamental component of the total space current is 0.275 x 116 = 31.9 milliamperes and the third-harmonic component is 0.185 x 116 = 21.4 milliamperes. For an amplification factor of 36 for the tube, the grid bias required for a

tripler operating with a plate voltage of 350 volts and an angle of plate current flow of 90 degrees is 42 volts. The angle of grid current flow is 78 degrees. The d-c and third-harmonic components of the grid current are 0.11 and 0.17, respectively, of the peak value of the grid current. The d-c grid current is 5.8 milliamperes and the crest value of the third-harmonic component of the grid current is 9 milliamperes. The d-c plate current is 16.8 - 5.8 = 11 milliamperes. The third-harmonic component of the plate current is 21.4 - 9 = 12.4 milliamperes.

The power input to the plate is 3.85 watts. The third-harmonic power developed in the plate tank is 1.86 watts and the plate dissipation is 2 watts. The tripler operates grounded grid and the driving power required is the total fundamental power in the cathode tank, or (0.0319 x 54)/2 = 0.862 watt. The grid-leak resistance required is 7,250 ohms, and the loaded plate tank impedance necessary for class C design conditions is 24,200 ohms.

Tank Circuit Design

The interelectrode capacitances of the 2C40 lighthouse triode are as follows:

Grid-to-plate capacitance = 1.3 micromicrofarads
Grid-to-cathode capacitance = 2.1 micromicrofarads
Plate-to-cathode capacitance = 0.02 micromicrofarad

Plate Tank. The plate tank for the 2C40 tripler is composed of a shorted section of flat-element transmission line. The tube is placed at the center of a half wavelength line shorted at both ends. The elements are 3 inches wide and spaced 3/8 inch apart.

Effectively the tank is two shorted quarter wave lines in parallel and the characteristic impedance of the lines is the same as if it were a shorted quarter wave line 6 inches wide. The characteristic impedance of a flat-element transmission line is[1]

$$Z_0 = 377\ s/w = 23.6 \text{ ohms for the plate line.}$$

The plate line is shortened by the grid-to-plate capacitance of the tube and the capacitance of the tuning condensers. Two variable tuning condensers are placed beside the tube. The maximum capacitance of one of these condensers is 0.7 micromicrofarad. For two tuning condensers in parallel at half their full capacitance, the total loading capacitance on the line will be 0.7 + 1.3 = 2 micromicrofarads. At 885 megacycles the capacitive reactance of 2 micromicrofarads is 90 ohms. To provide sufficient inductive reactance to resonate this capacitance at 885 megacycles a shorted transmission line with a characteristic impedance of 23.6 ohms must be 75.3 degrees or 0.209 wavelength long. The total distance between shorts for the plate tank is 5.58 inches.

Cathode Tank. The cathode tank of the 2C40 tripler is resonant at 295 megacycles and also made from a shorted section of flat-element transmission line. However, the cathode tank is a quarter wave line with the tube tapped down 2 inches. The line elements are 2 inches wide and spaced 3/8 inch apart which gives a characteristic impedance of 70.7 ohms. A single tuning condenser identical to those used for tuning the plate line is placed

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[1] B. E. Parker, "V. H. F. Tank Design," Radio and Television News, 43:8, January, 1950.

in the open end of the cathode line and the total loading capacitance is 0.35 + 2.1 = 2.45 micromicrofarads. A shorted transmission line with a characteristic impedance of 70.7 ohms must be 72.3 degrees or 0.201 wavelength long to resonate this capacitance at 295 megacycles. The cathode line is 8.05 inches long.

Plate Tank Efficiency

The attenuation of imperfectly conducting parallel flat plates has been calculated by Sarbacher and Edson.[1] The attenuation is

$$\alpha = \frac{1}{s}\sqrt{\frac{\pi f \mu_2 \varepsilon_1}{\sigma_2 \mu_1}} \text{ nepers per meter}$$

where s = the spacing between plates in meters
ε_1 = dielectric constant of the medium separating the conducting plates
μ_1 = permeability of the insulating medium
μ_2 = permeability of the conducting plates
σ_2 = conductivity of the conducting plates

For the silver-plated brass plate line with conducting plates spaced 3/8 inch and separated by air,

$s = 9.52 \times 10^{-3}$ meters
$\varepsilon_1 = 8.854 \times 10^{-12}$ farad per meter
$\mu_1 = \mu_2 = 1.257 \times 10^{-6}$ henry per meter
$\sigma_2 = 6.1 \times 10^{7}$ mhos per meter
$f = 885 \times 10^{6}$ cycles per second
$\alpha = 2.11 \times 10^{-3}$ nepers per meter

The input resistance of a shorted quarter wave line with a characteristic impedance of 23.6 ohms and an attenuation of 2.11×10^{-3}

[1] Robert I. Sarbacher and William A. Edson, Hyper and Ultrahigh Frequency Engineering (New York: John Wiley & Sons), 1947, p. 165.

nepers at a frequency of 885 megacycles is

$$R_{sc-\frac{\lambda}{4}} = \frac{4\ Z_o\ f}{\alpha\ c} = 1.32 \times 10^5 \text{ ohms}$$

The input resistance to the plate tank is less than the shunt resistance of a quarter wave line because the line is shortened by capacitive loading. From Plate XIII the shunt resistance of the plate tank is 0.94 $R_{sc-\frac{\lambda}{4}}$, or 1.24×10^5 ohms. The loaded shunt tank impedance required for class C design conditions is 24,200 ohms. The plate tank efficiency given by the equation $\frac{(R_{shunt} - R_{loaded})}{(R_{shunt})}$ is 80.6 per cent. The power output of the tripler is 0.806 x 1.86 = 1.5 watts.

2C40 POWER AMPLIFIER AT 885 MC

Class C Design

The angle of plate current flow for the power amplifier was chosen as 160 degrees. The plate supply voltage used was 350 volts. The ratings for the 2C40 have been listed in the 2C40 tripler design. The power amplifier operates with the same cathode emission and maximum grid voltage as the tripler.

On the assumption that α equals 3/2, the d-c and fundamental frequency components of the total space current are 0.25 and 0.425, respectively, of the peak space current. The d-c component of the total space current is 29 milliamperes and the crest value of the fundamental frequency component of the total space current is 48.3 milliamperes. Using an amplification factor of

36 for the tube and a plate supply voltage of 350 volts, the grid bias required for an angle of plate current flow of 160 degrees is 12.7 volts. The angle of grid current flow is 118 degrees. The d-c and fundamental frequency components of the grid current are 0.175 and 0.325 times the peak value of the grid current, or 9.26 and 17.2 milliamperes, respectively. The d-c plate current is 29 - 9.26 = 19.7 milliamperes, and the fundamental componont of plate current is 48.3 - 17.2 = 31.1 milliamperes crest value.

The power input to the plate is 6.9 watts. The power delivered to the plate tank is 4.65 watts and the plate dissipation is 2.25 watts. The grid driving power required is 0.596 watt and the grid-leak resistance is 1,370 ohms. The loaded tank impedance necessary for design conditions is 9,640 ohms.

Tank Circuit Design

Plate Tank. The plate tank is identical to the plate tank for the 2C40 tripler. The unloaded shunt tank impedance is 1.24×10^5 ohms, as previously calculated. The plate tank efficiency is 92 per cent and the power output of the amplifier is 4.28 watts.

Cathode Tank. The cathode tank construction is similar to the plate tanks of the 2C40 tripler and 2C40 power amplifier. The cathode tank is constructed of a section of flat-element transmission line shorted at both ends and the tube placed at the conter. The tuning condensers are placed beside the tube. The

elements of the line are 3 inches wide and spaced 3/8 inch apart. Consequently the cathode line has a characteristic impedance of 23.6 ohms which is the same as for the plate tank. The total loading capacitance on the cathode line is 0.7 + 2.1 = 2.8 micromicrofarads. Therefore the distance between the tubs and the short is 0.194 wavelength, or 2.59 inches. The total distance between shorts for the cathode line is 5.18 inches.